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SOVIET INSTRUMENTATION AND CONTROL TRANSLATION SERIES

Measurement Techniques

(The Soviet Journal *Izmeritel'naya Tekhnika* in English Translation)

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Measurement Techniques

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THE PROBLEM OF ANALYZING ERRORS IN MEASURING SYSTEMS

N. A. Chekhonadskii

Translated from Izmeritel'naya Tekhnika, No. 10, pp. 1-4, October, 1960

Modern remote measuring and automatic control systems, which for the sake of brevity, we shall henceforth call measuring systems, usually consist of fairly complex devices including, in a general case, several interconnected transducer links.

When measuring physical values which change continuously with time, these links are often subjected to various external disturbances: linear accelerations, variations in temperature and pressure, in supply voltages, etc. These disturbances may affect the work of the elements in converting their respective input signals, thus introducing certain errors in their output variables.

Since such conditions of operation of measuring systems are the most characteristic for them, it is necessary to base the problem of analyzing in a general form the errors in measuring systems precisely on such a case [1].

However, this problem has not as yet been satisfactorily solved for this, its most general form of application. The work known to us in this sphere [2, 3, 4] deals with the analysis of certain particular aspects of the problem. It should be noted that the random external disturbances affecting the work of the measuring systems can extend to a wider frequency range than the one covered by the system elements, and the measuring systems may have feedbacks.

We examine below the most simple and common case, when a measuring system consisting of several series-connected linear elements is affected by low frequency random external disturbances.

General considerations of the operation of a measuring system under the effect of external disturbances. Let us examine a system, intended for remote measurement of the pressure of liquids in mechanical equipment (Fig. 1), which consists of three elements – a potentiometer transducer (a), a measuring circuit (b) and a measuring instrument (c).

The pressure of the liquid is converted by the transducer into variations of the ohmic resistance in the measuring circuit, which also includes a source of direct current, whose value is varied by the circuit. The current is measured by the output instrument.

The characteristics of the measuring circuit elements can be considered in the first approximation to be linear, thus making it possible to represent the output variable of the first element for low variations of the input signal by the following expression:

$$R(t)=k_1P(t), \tag{1}$$

where k_1 is the transducer's transfer constant.

It can be easily shown that if the remaining elements of the measuring system are linear, its output variable can be represented by the following expression:

$$\alpha(t)=k_pP(t), \tag{2}$$

where $k_p = k_1k_2k_3$ is the total transfer ratio of the system.

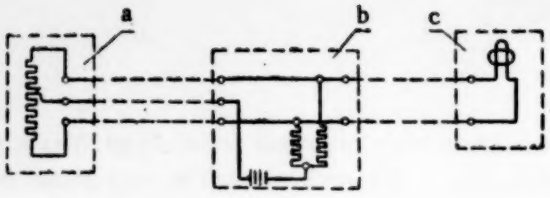

Fig. 1.

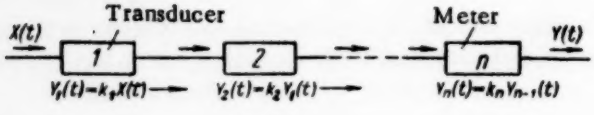

Fig. 2.

The above relationship, however, only holds for an ideal case, when each element of the system converts the input signal with complete accuracy and without any errors.

In fact, however, the following conditions will hold.

The measured pressure P(t) is converted by the transducer of the measuring system into resistance variations, with a certain static error, which is also a function of time:

$$R(t)=k_1P(t)+\delta_1(t), \tag{3}$$

where $\delta_1(t)$ is the transducer error.

The value of R(t) is then consecutively acted upon by the second and third elements of the measuring system, each element converting its input signal, which contains errors of the previous element, thus introducing at each stage certain errors due to conversion inaccuracies.

Similar effects will arise in the operation of any measuring system which consists of several elements connected in series.

Owing to the fact that the elements' conversion errors, which we shall call henceforth basic errors, are of the same order as the tolerances of the measuring systems, the problem of accounting for the conversion errors in the elements is of paramount importance in analyzing the accuracy of the system.

This, in effect, represents the basic difference between analyzing the errors of measuring systems and those of automatic control systems, whose conversion errors in elements are not taken into account.

Moreover, the measuring system elements can be affected by various slowly changing external disturbances.

Thus, the temperature of the air around the pressure transducer can vary with time, eventually leading to a certain change in the resistance of the variable portion of the transducer potentiometer, as compared with its value at the same pressure for a normal temperature. Such variations in the transducer resistance produce a static error, which we shall call henceforth an additional error.

Since the external disturbances affecting the measuring system elements can be represented as certain functions of time Z(t), the additional errors due to the system elements will also be functions of time.

Normally in measuring systems a linear relationship holds between the external disturbances and the additional errors caused by these disturbances. Hence an additional error arising in an element of the system under the effect of a jth disturbance can be represented by

$$\varepsilon_j(t)=\beta_j Z_j(t), \tag{4}$$

where $Z_j(t)$ is the functional relationship to time of the external disturbance which affects the given element of the measuring system; β_j is the "sensitivity" of the element to the external disturbance which affects it.

In the above relationship coefficient β_j expresses the relation between the jth external disturbance reacting on the given element and its additional error, which arises in that element and is due to the jth external disturbance.

If the elements of the measuring system under consideration (Fig. 1) are affected by the external disturbance, leading to the appearance of additional errors (for instance, ambient temperature variations), the output variable of the system can be represented by

$$\alpha(t)=k_pP(t)+\sum_{i=1}^{3}k_i^*\left[\delta_i(t)+\beta_iZ_i(t)\right], \tag{5}$$

where k_i^* is the transfer ratio of the measuring system's portion which consists of elements numbering from i + 1 to the last element of the measuring system.

The above expression can be made generally applicable to the case of a linear measuring system consisting of n elements connected in series (Fig. 2) and subjected to the action of m external disturbances:

$$Y(t)=k_p X(t)+\sum_{i=1}^{n} k_i^* [\delta_i(t)+\sum_{j=1}^{m}\beta_{ij} Z_{ij}(t)], \tag{6}$$

where X(t) is the input variable of the measuring system; Y(t) is the output variable of the measuring system.

Thus, when a linear measuring system is used and subjected to external disturbances, the measurement result will contain a total error due to the transfer to its output of the basic and additional errors of the elements.

Static characteristics of the total error of a measuring system. A measuring system's total error determined from (6) can be represented by

$$\delta(t)=Y(t)-k_p X(t). \tag{7}$$

Let us assume that the input of the measuring system repeatedly (N times) receives a variable $X_0(t)$ which is produced with great accuracy, exceeding that of the system under analysis. In such a case the measuring system will transform the input variable, producing at its output N values which may differ slightly from each other, despite the fact that one and the same function $X_0(t)$ was impressed several times on the input of the system.

This is due to the fact that the output variable of the measuring system is distorted by the total error, which consists of the basic and additional components.

It can be easily shown that these errors are, in a general case, random functions of time. Therefore, the total error of the measuring system can, in this case, only be evaluated statistically by working out from the N measured values of the output variable the expectation value of the total error:

$$M[\delta(t)]=M[Y(t)]-k_p X_0(t) \tag{8}$$

and the dispersion of this error:

$$D[\delta(t)]=D[Y(t)]. \tag{9}$$

It will be easily seen that the physical meaning of the total error expectation determined by (8) represents the systematic component of the measuring system error, and the total error dispersion represents the random component of the system error when it measures input variable X_0. When a variable is measured by a system under the effect of random external disturbances, these components are functions of time.

Analyzing the operation of a measuring system when it receives a random function of time. Let us examine the most general case, when the measuring system consists of n series-connected elements and the input variable X(t) is a random function of time.

In this case the actual value of the output variable can be found by means of appropriate transformations of the random functions contained in (6). In order to solve this problem let us use the methods developed in the theory of random functions [6].

It is known that in order to deal with random functions it is necessary to know their statistical characteristics, including their expectation values and correlation functions of the random functions being summed up, as well as correlation functions of the relation between all the random functions taken in pairs.

Let us assume that all these characteristics are known. Then the expectation value of the measuring system output function will be represented by the relation

$$M[Y(t)]=k_p M[X(t)]+$$
$$\sum_{i=1}^{n} k_i^* \left\{ M[\delta_i(t)] + \sum_{j=1}^{m}\beta_{ij} M[Z_{ij}(t)] \right\}, \tag{10}$$

where $M[X(t)]$, $M[\delta_i(t)]$, $M[Z_{ij}(t)]$ are the expectation values of the input function, the ith element basic error and the jth external disturbance which affects the ith element.

The correlation function of the measuring system output function can be represented as

$$K[Y(t,t')]=k_p^2 K[X(t,t')]+\sum_{i=1}^{n} k_i^{**} K[\delta_i(t,t')]+\sum_{i=1}^{n} k_i^{**}\left\{\sum_{j=1}^{m}\beta_{ij}^2 K[Z_{ij}(t,t')]\right\}+\overline{R}_{x\delta}(t,t')+\overline{R}_{xz}(t,t')+$$
$$+\overline{R}_{\delta z}(t,t')+\overline{R}_{\delta}(t,t')+\overline{R}_{z}(t,t'), \qquad (11)$$

where $K[X(t, t')]$, $K[\delta_i(t, t')]$, $K[Z_{ij}(t, t')]$ are the correlation functions of the measuring system input function, the ith element basic error and the jth external disturbance affecting the ith element; $\overline{R}_{x\delta}(t, t')$ is the total correlation function of the measuring system input function for the basic errors of each of the n elements of the system; $\overline{R}_{xz}(t, t')$ is the correlation function of the relation of the measuring system input function to all the external disturbances; $\overline{R}_{\delta z}(t, t')$ is the total correlation function of the relation of the measuring system elements' basic errors to all the external disturbances; $\overline{R}_{\delta}(t, t')$ is the total correlation function of the inter-relation of all the system elements' basic errors; $\overline{R}_{z}(t, t')$ is the total correlation function of the inter-relations of external disturbances affecting all the measuring system elements.

In (11) the total correlation function of interrelations $\overline{R}_{x\delta}(t, t')$ is understood to represent the sum of the correlation functions of the relation of random function $X(t)$ to each of random functions $\delta_i(t)$, a sum determined according to the well-known rules outlined in [6]. This definition also refers to the remaining total correlation functions contained in the above expression.

The expressions (10) and (11) thus obtained determine in a general form the output variable of the measuring system when its input receives a random function of time, and the system is subject to random external disturbances.

These expressions can be used to analyze in a general form the errors of a measuring system. For this purpose it is, however, necessary to know, as previously stated, the statistical characteristics of the analyzed errors, including their expectation values and dispersions.

The latter characteristic can be obtained if in (11) we assume that $t = t'$.

After the required substitution we obtain

$$D[Y(t)]=k_p^2 D[X(t)]+\sum_{i=1}^{n} k_i^{**}\left\{D[\delta_i(t)]+\sum_{j=1}^{m}\beta_{ij}^2 D[Z_{ij}(t)]\right\}+R_{x\delta}^{*}(t)+R_{xz}^{*}(t)+R_{\delta z}^{*}(t)+R_{\delta}^{*}(t)+R_{z}^{*}(t), \qquad (12)$$

where $R_{x\delta}^{*}(t)$, $R_{xz}^{*}(t)$, $R_{\delta z}^{*}(t)$, $R_{\delta}^{*}(t)$, $R_{z}^{*}(t)$ are total interrelation functions of the corresponding random functions.

As the result of the analysis of (10) and (12) it is possible to draw the following conclusions:

a) The expectation value of the measuring system's output function contains a systematic error due to the effect of random external disturbances on the system. The value of this error depends on the expectation values of each of the external disturbances, on the sensitivity of the system elements to external disturbances and on the transfer ratios of the elements.

If the sensitivity of the elements to external disturbances is such that for all coefficients β the relation $\beta > 0$ holds, the expectation value of the output function, i.e., the measurement result, will be distorted even more by a systematic error than by the presence of external disturbances. With a relation of $\beta < 0$ a compensation of systematic errors is possible.

b) The dispersion of the measuring system's output function depends on the dispersions of the external disturbances, the sensitivity of elements to the external dispersions, the elements' transfer ratios, and the interrelation functions of all the random functions taken in pairs which determine the measuring system's output function.

The negative interrelations between the random functions which determine the measuring system's output function provide a reduction in the dispersion of the measuring system's output function. Thus, in this instance, a compensation of the measuring system's random errors can be achieved.

Expressions (10) and (12), which determine in a general form the output variable of the linear measuring system subjected to the effect of random external disturbances, can be used for analyzing errors in measuring systems for various particular cases of their operation.

SUMMARY

In the operation of a linear measuring system with its elements subjected to various random external disturbances which produce in the elements additional static errors, transmitted to the output of the system together with the basic errors of the elements, a total static error is produced at the output. The latter error is the one which determines the accuracy of the measuring system under given conditions.

If in experimental work the required variable is measured several times by means of a measuring system, whose elements are subjected to various random external disturbances, a statistical analysis of the measurement results may produce a certain compensation of the static measurement errors of this variable.

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NEW PRINCIPLES FOR DETERMINING COMPULSION IN STATE CHECKING OF MEASURING INSTRUMENTS**

V. D. Alesin

Translated from Izmeritel'naya Tekhnika, No. 10, pp. 5-7, October, 1960

The July Plenum of the Central Committee of the Soviet Union Communist Party has outlined important measures for speeding up technical progress and the growth of labor productivity, improving production organization, raising the level of commercial work, and improving the control of industry and transport.

Since the work of the state inspection laboratories for measuring equipment of the Committee of Standards, Measures, and Measuring Instruments is closely connected with the problems discussed by the Plenum, it is expedient to discuss certain aspects of the inspection work of the Committee's local agencies, in order further to improve the work of state inspection laboratories for measuring equipment. N. S. Khrushchev stated in his report to the Twenty-first Congress that at present we require above all concrete thinking, purposeful action and a profound knowledge of our subject.

The Committee's local agencies have been continuously working for dozens of years with the object of providing a uniform, accurate and correct application of measures and measuring instruments in our national economy. The major part of this activity of the Committee and its local agencies consists of inspection work aimed at providing a constantly efficient and accurate operation of measuring instruments in our national economy.

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*See English translation.

**As a contribution to a discussion.

However, despite the very large amount of work carried out in this connection by the Committee's state inspection laboratories, the present state of measuring equipment in many plants and organizations cannot be considered satisfactory.

At the present it is the type of the instrument which determines, irrespective of the sphere of its application, whether the instrument is subject to compulsory testing (stamping) or not. This condition leads to many shortcomings in the work of the Committee's local laboratories.

This is made even worse by the fact that at present the work of these laboratories is evaluated by the amount of money they collect for testing measuring instruments according to plan. Such control methods often encourage the Committee's local agencies to test as many instruments as possible, without considering whether this will provide efficient measuring equipment in any given plant, especially for the main production processes (control of the basic technological operations, checking finished production, testing various materials and parts, etc). Since the necessity of compulsory testing of instruments is determined by their types the Committee's local agencies do not check or supervise the operation of such modern measuring equipment as automatic sorting apparatus, various transducers, measuring means in automatic receiving and control checking (production control) and other measuring equipment, which plays an important and vital part in production automation, in the assimilation of technological processes and new equipment by our industry.

Determining the necessity for compulsory state checking of measures and instruments by their types also leads to a tendency of the Committee's local agencies for testing all the instruments used in our national economy (without possessing the necessary list of instruments or data on their condition in any given economic region). This creates a situation in which the amount of work becomes unknown and hence its distribution is mainly determined by the desire to collect the sum of money set by the plan for testing any instruments (without, of course, taking account of the sphere of their application).

It would be more efficient and far better if the inspection work of the local laboratories of the Committee were based on the following fundamental principles and designed to satisfy the basic requirements of our national economy in connection with technical progress, raising labor productivity, increasing the volume of production, improving its quality and lowering the costs of production.

1. The inspection of measures and measuring instruments used in our national economy should be differentiated with respect to the sphere of their application and distributed between the Committee's state agencies and service departments of various plants and organizations, which should carry out state testing (by the Committee's agencies) and service testing (by other organizations). Both types of tests should be compulsory and periodic.

2. The list of measuring instruments subject to compulsory state testing and stamping should depend on the sphere of their application and be determined, with an indication of the department and sphere of their application, separately for each plant or organization, by compiling special lists which should include the measures, measuring instruments, measuring and special automatic devices and other measuring equipment subject to checking and stamping by the Committee agencies. The checking periods for these instruments should be set according to the conditions of their use.

3. Compulsory testing and stamping by the Committee's agencies should be carried out on the measuring equipment used in the following spheres (branches) of our national economy: instruments used for checking or calibrating other measures or instruments, i.e., used as reference standards; instruments used for recording the output or consumption of various types of energy (electrical, gas, steam, water, etc.), or materials or products (oil, alcohol, milk, etc.); instruments used in trade or other operations when payment is made according to the values measured; instruments used in controlling important technological processes in production; instruments used in checking finished articles or production by the Technical Control Divisions; instruments used for checking deliveries to the state of agricultural products or transactions with the collective farms; instruments used for measuring important freight and materials transported by road, railway, river, sea, air or other means; instruments used in the operation of important installations or assemblies; in the protection of the workers' health; in safety devices; in state testing of installations, machinery and materials; in expert advice or arbitrations; in acceptance or delivery tests carried out in plants and other organizations.

Other spheres of application of instruments tested by the Committee's agencies can also be established according to the specific conditions prevailing in plants and organizations.

4. All other measures and measuring instruments which are not included in the list of those subject to compulsory state testing by the Committee's agencies should be tested by the plants or organizations in their own laboratories or by other organizations (including the Committee's laboratories, if required) in the manner determined by the owners.

5. The dates (intervals between) of state testing are established by the Committee of Standards, Measures and Measuring Instruments, and the dates (intervals between) of service testing are determined by those responsible for the organizations where the tests are carried out.

6. The state testing of instruments is carried out according to the technical instructions issued by the Committee of Standards, Measures and Measuring Instruments, and the service testing can be carried out either according to the Committee's instructions, or according to the methods and by means of the equipment specified by the plants and organizations carrying out the test.

7. Instruments which, as a rule, are used as reference standards for checking the output of energy, products or materials, in trade, in safety devices and for the protection of the workers' health, are subject to compulsory state checking and stamping before leaving the factory in which they are produced.

Concrete lists of such instruments should be compiled by the local agencies of the Committee for each instrument-making plant.

8. The instruments which have passed government tests are stamped (or supplied with a certificate) by the Committee's agencies. The method of certifying instruments which have passed service testing should be established by the plants and organizations themselves.

9. The Committee's agencies, which are combined in a single state service of measures and measuring instruments of the USSR, in exercising their right of checking, inspecting and investigating the conditions, utilization, manufacture, repair, storing and sales of measuring equipment in plants and organizations, should concentrate not only on finding defects, but mainly on a thorough study of the equipment and on rendering practical assistance to any given plant or organization in developing, improving and assimilating new measuring equipment.

10. Measures and measuring instruments repaired in commercial shops may be subjected, in order to check the quality of repairs, to compulsory state testing irrespective of the sphere of application of these instruments.

11. The sum of money obtained by the Committee's agencies as fees for state testing of instruments should not be considered as a measure of their activity.

12. Neither should state testing of instruments by the Committee's laboratories constitute the object of their work, but serve as a means for implementing their main task of providing a uniform and accurate measuring equipment, in rendering practical help to plants and enterprises in developing their means of measurement, in improving measuring apparatus and providing scientific and technical assistance for assimilating new measuring techniques.

In particular it is desirable that the Committee's laboratories should adjust and tune complex measuring devices in situ, test measuring equipment on request of plants and organizations, send teams of specialists for organizing in the plants efficient inspection, utilization, testing and adjusting of measuring equipment, and organize special courses for improving the qualifications of the industrial laboratory personnel.

13. The organization of work and administration of the Committee's local laboratories should direct all their activity and initiative to studying the requirements of the economy of their own region and rendering practical assistance in developing this region and accelerating its technical progress in the sphere of measurement techniques.

One of the most important aspects in the work of the Committee's local laboratories should be the development and further improvement of their laboratory base (including the establishment of repair and adjusting workshops and design groups attached to the Committee's laboratories), in order to create material and technical means for satisfying fully the requirements of our national economy and anticipating them in all the spheres of measurement, both with respect to accuracy and range, taking into consideration not only the present level of the measurement technique of their respective economic regions, but also the future development and further technical progress in their regions. Thus, the laboratories of the Committee should be on a higher plane than

any industrial laboratory with respect to equipment, qualified personnel, accuracy and technique of measurements, and should be real metrological and organizational centers for measurement technique of their economic region.

If the sphere of application of the instruments is taken into consideration when their testing is organized, the Committee's laboratories will have to develop their work on the basis of the concrete requirements of the economy of their region and deal in the first place with the spheres of measurement of greatest interest to the local plants and most important for the development of the local economy.

The proposed method of organizing the work provides a solid basis for planning the operations of the Committee's local laboratories, and hence for evaluating their activity and organizing socialist competition.

The reorganization of the work of the Committee's local agencies in the manner suggested above will provide them with creative organizational and technical work directly in the plants in close collaboration with the factory workers, and will not only radically improve the conditions of measurement equipment in the factories and organizations of their economic region, but also create better conditions for improving the qualifications of the Committee's laboratory personnel, will widen their outlook and enrich their experience with valuable practical knowledge.

The activity of the laboratory workers will be concentrated on fulfilling the main tasks of the Seven-Year Economic Plan and on assisting in every possible way the technical development of their region.

Editorial Note: The Editorial Board requests the readers of this journal to comment on the points raised in this article.

TECHNICAL AND ECONOMIC PROBLEMS OF CHECKING COMPONENTS IN ENGINEERING

A. M. Vedmidskii

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Automatic checking of the dimensional parameters of components plays an important part in full production automation in engineering and instrument-making. The effective solution of the technical problems in checking components requires a special technical and economic analysis.

The main technical-economic index which characterizes the efficiency of testing operations under production conditions in engineering or instrument-making is the cost of checking components. The most important factors making up the cost of checking machines and instruments consist of the required measuring accuracy corresponding to the tolerances in the deviation of parameters, and the choice and assignment of the measuring equipment (taking into consideration the reliability of checking, i.e., preventing any defective components being assembled); the time required for checking dimensional parameters; and the expenses of running and repairing the testing equipment.

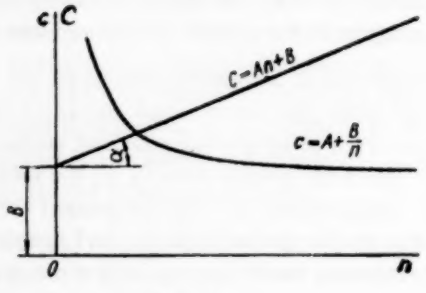

Fig. 1.

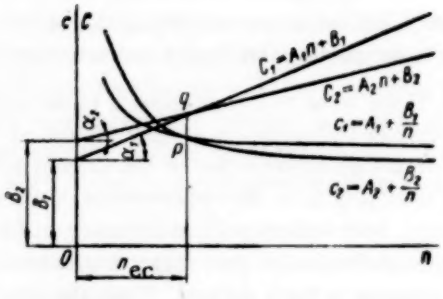

Fig. 2.

When selecting measuring equipment it is necessary to take into account their initial cost.

Components can be checked in various ways, according to the selected measuring apparatus and its relation to the main production equipment. The most expedient production control methods in the economic and metrological respects are those which provide the smallest possible expenditure on checking without raising the cost of producing and assembling components. The selection of such a method for mass or conveyor-belt production is mainly determined by the production volume. The most economical control method can be selected on the basis of the following considerations.

The expenditure related to the checking of components can be divided into two basic groups:

1) unit expenditure, i.e., expenditure repeated for each component;

2) total initial expenditure incurred but once, in preparing production for all the components to be tested during the manufacturing process, or incurred periodically for each batch of components.

Thus, the cost of checking of one component is expressed by the formula

$$c = A + \frac{B}{n},$$

where A is the sum of current expenses for checking one component; B is the sum of the initial total expenditures; n is the number of checked components.

It will be seen from the above formula, which represents an equation of a hyperbola (Fig. 1), that for a given method the cost of checking one component depends on the number of controlled components. The cost of checking falls with a rising number of components.

In order to compare the cost of checking a component by means of two different methods (for instance, by means of mechanized and automatic control) it is possible to plot on the same graph two curves which correspond to the two equations representing each method (Fig. 2):

$$c_1 = A_1 + \frac{B_1}{n};$$

$$c_2 = A_2 + \frac{B_2}{n},$$

where A_1 and A_2 are the total current expenditures for checking one component by means of two different methods; B_1 and B_2 are the total initial expenditures.

In a general case, when the curves on the graph cross each other, for instance at point p, the value of the abscissa at the crossing point corresponds to the number of components n_{ec} for which the cost of checking is the same in both methods. If the actual number of checked components is smaller than n_{ec}, the first method will provide a lower cost of checking. If the number of components is larger than n_{ec}, the second method will be more economical.

The number of components for which the cost of checking will be the same in both methods can be determined directly from the formula

$$n_{ec} = \frac{B_2 - B_1}{A_1 - A_2},$$

obtained from the previous two equations.

For an economic analysis of the various checking methods it is possible to compare the cost of checking the total number of components. In fact if the cost of checking one component is c the total cost C of checking n components amounts to

$$C=cn$$
or $$C=An+B,$$

i.e., it is represented by a linear equation.

The total cost of checking a batch of components with respect to their number in each batch for two different control methods can be represented graphically by two sloping straight lines (Fig. 2), plotted from the two equations

$$C_1=A_1n+B_1$$
and $$C_2=A_2n+B_2.$$

It will be seen from Fig. 2 that $A_1 > A_2$, but $B_2 > B_1$. However, despite the larger initial expenditure on measuring equipment (automatic checking), the cost of checking is lower for the second method, owing to a smaller expenditure on wages, providing the number of checked components is larger than n_{ec} (the actual point of intersection is to the right of point g).

When one of the methods has the values of both A and B larger or smaller than those of another method, the numerical value of n_{ec} will become negative (with a minus sign), i.e., the intersection of the inclined lines will be to the left of the Y axis. In practice it means that one of the methods is more economical than the other over the whole range of O–n. Moreover, the difference in the cost of checking will change according to the number of checked components, as shown by the different values of the angle of slope α.

The same method can be used for comparing the cost of checking components by means of several methods. For this purpose it is only necessary to derive an equation for each method and plot it on a common graph, and then find according to the required number of tested components which of the methods is the most economical.

The equations and graphs for analyzing the cost of checking components depend, of course, on the specific production conditions. In order to raise the accuracy of the economic analysis it is advisable to estimate within determined limits the number of checked components.

The methods of determining the basic elements in the cost of checking components are closely connected with an expedient distribution of expenditure required for the checking operations. The nature and size of the expenditure vary with the operational and metrological characteristics of the testing equipment, which can be classified from the point of view of their technical and economic analysis into universal, special, mechanical and automatic means of control. Their common characteristic index consists of the previously adopted distribution of expenditure.

The current expenditure on testing one component can be represented by the formula

$$A=\sum(l_\partial+p_\partial),$$

where l_∂ is the wage paid for each checking operation; p_∂ is the overhead expenditure.

The wages paid for each checking operation are proportional to the time spent on it and are determined from the formula

$$l_\partial=\frac{q_{hr}\,T_1}{60}=\frac{q_{hr}}{n_{hr}},$$

where q_{hr} is the hourly wage; T_1 is the time spent by the controller (workman) on checking one component, in minutes; n_{hr} is the number of components checked in one hour.

The overhead expenses represent the expenditure per checking operation on depreciation, repairs and maintenance of buildings, testing equipment, devices, instruments and tools of general use, the cost of electrical power, the salaries of the administrative and technical personnel (of the Technical Control Division), etc. For economic calculations overhead expenditure should be estimated on the basis of similar workshops and test departments. Moreover, in order to prevent considerable inaccuracy, it is not advisable to take the average relative overhead expenditure as a basis for calculation.

The initial expenditure on testing equipment can be classified into: 1) expenditure connected with the preparation, starting and adjusting of automatic equipment, instruments, devices and measuring instruments; 2) expenditure on purchasing special automatic equipment, instruments, devices and measuring instruments.

The first item of expenditure connected with preparatory and finishing operations consists of the wages of workmen (controllers) or adjusters, paid for preparing and adjusting the testing equipment for a given batch of similar components, as well as for the dismantling and storing of this equipment after testing. The total expenditure connected with preparatory and finishing operations consists of wages and overhead expenses per batch of components, multiplied by the number of batches.

The second item of initial expenses consists of the expenditure on purchasing, manufacture, repair and periodic adjustment of special testing apparatus which is used only for checking a given batch of components.

The total amount of the initial expenditure can be determined from the formula

$$B=\sum (l_a+p_o)\, n_o+\sum C_m\left(1+\frac{q_m}{100}\right) n_i ,$$

where l_a is the wage paid for one adjustment; p_o is the overhead expenditure on the wages paid for adjusting; n_o is the overhead expenditure on checking operations; C_m is the cost of special measuring instruments (automatic devices), apparatus, etc.; q_m is the overhead expenditure on the maintenance and repairs of special measuring equipment in %; n_i is the number of special measuring instruments and tools of a given type used in each checking operation.

The wages paid for each adjustment correspond to the time spent on the preparatory and finishing operations which, in turn, depend on the complexity and accuracy of the measuring automatic devices, instruments and tools. For checking a large number of components readjustment of the apparatus is also required (for instance, in automatically checking and sorting components into groups). In such cases the additional workers' wages and expenditure on readjustments should all be taken into account.

The initial cost of a special measuring instrument or automatic device (C_m) consists of its cost proper, and the expenditure on additional attachments and on acquiring the knowledge how to use it.

The total life of a measuring instrument or automatic device depends on metrological and technological factors. For instance, the life of an instrument depends on the value and position of its actual error with respect to the specified tolerances. If the instrument's transducer has to operate in an area covered with abrasive particles its rate of wear is greatly increased.

Thus, the life and hence, depreciation expenditure, varies from instrument to instrument. If the instrument has a long life only a part of its total cost C_m should be included for determining the value of B. However, the full cost of an instrument with a short life or a highly specialized design should be applied to the first batch of tested components.

A correct solution of the technical and economic problems in selecting and adapting modern equipment for checking components helps to provide an efficient development of total production automation in plants.

CLASSIFICATION OF PHYSICAL QUANTITIES BY THEIR KINDS

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In [1, 2] the necessity of supplementing the theory of dimensions by numerical quantities was pointed out. This problem is treated below in a considerably modified and more general manner.

The theory of dimensions deals with quantities of two categories: dimensional quantities expressed by denominated numbers, and dimensionless quantities expressed by abstract numbers.

We consider that all physical quantities should be classified by the kind of numbers and properties of measurement units into three instead of two categories, consisting of dimensional quantities, numerical enumerated quantities, and relative numerical quantities.

Dimensional quantities. The properties of dimensional quantities are well known and therefore we shall deal with them briefly. It is characteristic of a dimensional quantity that it is expressed by a number (measured) by comparing it with a unit of measurement, which represents a specimen of that quantity. The number which represents after measurement the dimensional value of the quantity reflects the properties of two specimen quantities, the one being measured and an auxiliary quantity (unit of measurement). This number is denominated; it includes an abstract numerical part and a designation consisting of the name of the measuring unit. It is convenient to denote the dimensional quantities (denominated numbers) as

$$v = |v| \operatorname{Nom}[v]. \tag{1}$$

Here $|v|$ is the numerical value, Nom (nomen=name) is the name of the measuring unit, $\underline{v}$ is the denominated value.

The measurement unit itself can be written in the form

$$[v] = 1 \operatorname{Nom}[v]. \tag{2}$$

This denominated number 1 Nom [v]–denominated unit–can be called the dimension of the measurement unit. In the present theory of dimensions the term "dimension of a unit" has a rather vague meaning; we always understand the dimensions of a unit to be a number which expresses the unit of measurement, in this case the denominated number 1 Nom [v].

Operations with denominated numbers consist of operations on their abstract numerical parts $|v|$ and operations on the designations. Moreover, operations on the designations mean that as a result of them new designations for derived units are always obtained with a single-valued definition in terms of this operation.

For instance, let us divide the denominated number expressing displacement by that of time: $s = |s| \operatorname{Nom}[s]$ and $t = |t| \operatorname{Nom}[t]$:

$$v = \frac{s}{t} = \frac{|s| \operatorname{Nom}[s]}{|t| \operatorname{Nom}[t]} = \left(\frac{|s|}{|t|}\right)\left(\frac{\operatorname{Nom}[s]}{\operatorname{Nom}[t]}\right). \tag{3}$$

As the result of this operation Nom [s]/Nom [t] must be interpreted as the designation of a definite unit of velocity Nom [v]. The quantity [v] = 1 Nom [v] has a single-valued definition as the velocity of a uniform motion in which the body covers a distance equal to 1 Nom [s] in time t = 1 Nom [t].

Enumerating numerical values. We shall define as numerical all physical quantities which by their very definition represent a number. Contrary to dimensional quantities a number which represents a numerical unit has a meaning irrespective of the comparison of two specimens and characterizes a single specimen. We shall define as enumerated those quantities which represent the number of objects in an aggregate, for instance, the number of vibrations, the number of molecules, the number of microstates, etc.

A similar part is played in the theory of dimensions by certain other values derived from the fundamental above-mentioned values by means of mathematical operations, for instance, such as entropy considered as the logarithm of the number of microstates.

The measurement units of above quantities consist of a special kind of measuring units, "specimen-units." Specimen-units are not simply abstract numbers; they retain a certain kind of individuality and can be expressed by numbers with subscripts such as 1_q. The specimen-unit for quantity $\underline{q}$ is 1_q. Let q = N = the number of molecules. It is important to understand that the symbol 1_N does not mean "1 molecule." Here the concept "molecule" is neither a physical quantity, nor the name of a unit; the quantity consists only of a number of molecules; thus 1_N is also a number (unit) with a reference to the individuality of the quantity under consideration (a specimen-unit for the number of molecules).

The value of the above quantity is expressed by a number of a special kind:

$$q = |q|\, 1_q, \tag{4}$$

which can be called an improper abstract number (where 1_q is an improper numerical unit, and $|q|$ is a completely abstract number).

The measurement unit indexes in enumerated quantities have an important function.

One of their most important functions is revealed in multiplying or dividing by an enumerated quantity which always leads to a new quantity with a qualitatively different individuality. Thus, energy ϵ and the energy referred to one molecule ϵ/N which is a measure of temperature, undoubtedly express different physical concepts. Similarly, time $\underline{t}$ and a period of vibrations T = t/F are quantities of a different individuality.

Let us now examine dimensional formulas with enumerated quantities. As an example let us take quantities T and ν, the period and frequency; they are expressed by means of an enumerated quantity, the number of oscillations F, in the following manner:

$$T = \frac{t}{F}; \tag{5}$$

$$\nu = \frac{F}{t}. \tag{6}$$

The dimensional formulas, i.e., the formulas which express the dimension of the derived units in terms of the dimensions of the original quantities, will have the form

$$[T] = [t]1_F^{-1} = \left[t 1_F^{-1}\right]; \tag{7}$$

$$[\nu] = \left[1_F t^{-1}\right]. \tag{8}$$

It will be seen that in this case the dimension is represented by a combination of denominated units and improper unit numbers with subscripts.

Enumerated quantities play an important part in molecular physics. The number of microstates P, entropy $\sigma = \log P$, the number of molecules, the number of degrees of freedom are all enumerated quantities. In a general case the temperature is represented by

$$\theta = \left(\frac{\partial \varepsilon}{\partial \sigma}\right)_v, \tag{9}$$

and has according to the above reasoning the dimension of

$$[\theta] = \frac{1\,\mathrm{Nom}\,[\varepsilon]}{1_\sigma}. \tag{10}$$

In a particular case expressed in CGS units

$$[\Theta] = \frac{1\ \text{erg}}{1_\sigma}, \tag{11}$$

whereas an ordinary degree is equal to

$$1^\circ = 1.38 \cdot 10^{-16}\ \text{erg}/1_\sigma . \tag{12}$$

Thermal capacity, determined by the ratio $\Delta\Theta/m\Delta t$, has the dimension of

$$[c] = \frac{[\epsilon]\ 1_\sigma}{[m]\ [\epsilon^1]} = \frac{1_\sigma}{[m]}. \tag{13}$$

The presence in the dimensional formula of the unit 1_σ completely corresponds to the physical meaning of thermal capacity, which is normally determined by the number of the degrees of freedom corresponding to a unit of mass. Let us note that the number of gram-molecules (moles) is not an enumerated quantity (despite its name), but represents a mass of special units. Hence dimensions of molecular quantities referred to a molecule always contain in their numerator a term "1 mole," instead of a term 1_q. The enumerated quantities have an important inherent property: various enumerated quantities can be equated to each other. Such expressions as "number of waves = number of oscillations" are completely legitimate and in many instances have a direct physical meaning (each oscillation produces one wave). The possibility of equating various enumerated quantities demonstrates their numerical character, in a certain sense, their "dimensionless" nature.

If we denote the number of waves by N, and the number of oscillations by F, the above equality, which was given as an example, can be written in greater detail as

$$|N|\,1_N = |F|\,1_F. \tag{14}$$

It will be seen from the above that physical equations which contain enumerated quantities can have different dimensions with respect to unit 1_q in their left and right hand parts.

In molecular physics equations with enumerated quantities and different dimensions in their right and left hand parts occur frequently. As an example of such an equation we can cite the relation between the mean kinetic energy of a molecule and temperature:

$$\bar{\epsilon} = \frac{3}{2}\Theta. \tag{15}$$

Here the dimension of the left hand side is $\frac{[\epsilon]}{1n}$, and of the right hand side is $\frac{[\epsilon]}{1_\sigma}$. It appears that when the dimensions of the left and right hand sides of an equation coincide only with an accuracy of 1_q the following rule will hold: the number of operations on the measurement units (their dimensions) for both parts of the equation must either be equal or differ by an even number.

Relative numerical quantities. In this category we include quantities which are determined by a ratio of two similar dimensional or enumerated quantities.

As examples of relative quantities we can cite the atomic weight, efficiency, permittivity and permeability. An important relative quantity is related to angles. The angle itself is a dimensional quantity, which is demonstrated by the possibility of arbitrarily using a denominated measurement unit and measuring the angle by its superposition. Angles are not related to other quantities by expressions which could be considered as their definition. Therefore angles must be classified as fundamental quantities similar to the length, time and force or mass.

The angle occupies a special place in dimensional quantities, because each angle has a single-value representation by means of a relative quantity. According to the law of proportionality between the central angle and its arc the following equality holds:

$$\frac{\alpha}{\alpha_0} = \frac{s}{r}, \tag{16}$$

where α is the given angle, α_0 is an angle whose arc is equal to the radius. The ratio $\frac{s}{r}$ can be called the numerical equivalent of an angle. The established practice of identifying the ratio $\frac{s}{r}$ with the angle and hence considering it as a dimensionless quantity will not stand up to critical examination and can only be retained nominally.

The trigonometrical functions of an angle are in fact functions of its numerical equivalent; they are also relative quantities derived from certain ratios in a triangle.

The measurement unit of relative quantities is a simple numerical unit without an individual subscript, and its meaning can be expressed by the term "once-unit." Relative units always function as "coefficients of amplification or attenuation." Multiplication or division by a relative unit does not lead to the formation of new concepts. The division of energy by the number of molecules (as enumerated quantity) leads to the formation of a new quantity $\frac{\epsilon}{N}$ with new properties and a new measurement unit, but the multiplication of the magnetic field strength H by μ or the division of a side of a right-angled triangle by the sine of the numerical equivalent of an angle does not change any concepts; the field strength remains a field strength and the length a length.

It follows from the above that the relative quantities and their measuring units are expressed by completely abstract numbers. Hence the "once-unit" does not figure in any of the dimensional formulas, but it is incorporated with the numerical values of the dimensional and enumerated quantities. For instance, the dimension of a charge from Coulomb's law is equal to $[l f^{\frac{1}{2}}]$ in which ϵ (the permittivity) does not figure.

SUMMARY

The division of all the physical quantities into the three categories specified in this article provides a deeper understanding of physical laws and makes it possible to establish a more consistent symbolism in the theory of dimensions.

A correct conception of measurement units of relative and enumerated quantities may play an important part in establishing a convenient, practical system of units.

Editorial Note: The editorial board considers it advisable for the readers of this journal to express their opinion on the essential problems raised in this article.

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LINEAR MEASUREMENTS

REPRODUCTION AND TRANSMISSION OF THE VALUE OF AN ANGLE IN REFERENCE MEASUREMENTS

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The rising requirements in the accuracy of reproducing and transmitting the value of a given angle and the increased production of precision angle-measuring instruments made it necessary to develop new means, methods and specifications for angle measurements, in order to insure their uniformity. According to these requirements the Sverdlovsk branch of the VNIIM (All-Union Scientific Research Institute of Metrology) developed a new test circuit; the VNIIK (All-Union Scientific Research Institute of the Committee of Standards, Measures and Measuring Instruments) compiled and issued an instruction for checking goniometers [1] and developed a horizontal equipment for measuring the angles of 1st grade polyhedral prisms [2]; and the VNIIM produced an equipment which provides a reference method of reproducing the value of a plane angle [3].

The reference angle measuring method, which is basic to the whole measuring system, consists in determining the angles of polyhedral prisms by calibrating them with two independent autocollimators. These instruments are placed in such a manner that the optical axes of their objectives are perpendicular to the reflecting surfaces of the prism, and the angle between the collimators is equal to the nominal value of the central angle between the perpendiculars to the sides of the prism.

The above method is one of the simplest and most accurate ways of reproducing and transmitting discrete values of an angle.

This method was applied in the equipment made by VNIIM which consists of two precision autocollimators with telephoto lenses (focal length of the objectives is 1000 mm) and three fused quartz reference prisms (two with 36, and the other one with 24 sides).

The equipment can be used for absolute measurements (calibration) and relative measurements (comparing angles of unknown prisms and angle-gauges with those of one of the reference prisms). By means of comparison methods it is possible to transmit the value of the angles of calibrated reference prisms to the lower-grade standards of the reference system; for this purpose the angles of the gauges being compared must be equal to or multiples of 10 or 15° (according to the values of the central angles between the sides of a reference prism).

The prisms being calibrated or compared are placed on a table, fixed to the axle of an optical dividing head, which serves as a turning device. All the instruments are placed on a solid cast iron plate.

The VNIIM installation differs from those used abroad by its improved measures and instruments. Thus the VNIIM autocollimators use long focal length telephoto lenses, which are designed for use in telescopic instruments with a small field of vision and make it possible to construct the instrument with a short mechanical length (less than 600 mm, including the eyepiece micrometers). Such a design made it possible to construct instruments which combine a high precision with easy handling. The autocollimators are supplied with screw eyepiece micrometers and optical micrometers calibrated in 0.25". Tests of the instruments have shown that the most important error was the error of reading, which arises when the crosshairs of the screw eyepiece micrometer are adjusted to coincide with the image of the autocollimator graduation. Under the most favorable conditions the quadratic mean error of reading was equal to $\sigma = 0.16"$ and was obtained by means of several adjustments to the image of the autocollimator graduation in the shape of light cross lines; the total magnification of the system was then

equal to 46.4 diameters. The errors of the micrometer screws were checked on a contact interferometer [4] and by means of three precision optical wedges, which were fixed to a flat plate (a device used for calibrating graduated thimbles of eyepiece micrometers), and were found in the operating range to be negligibly small compared with the reading error.

The most complicated part in producing this equipment was the making of reference measuring prisms. The difficulty arose in the necessity to combine the requirements of a high precision in angles and great stability of the prism. The development of a special technique, which obviated the accumulation of any appreciable error in the angles between the sides of the prism when they were being ground, made it possible to maintain the strict tolerances in a monolithic construction [5]. All the prisms were made from fused quartz disks of a diameter of 125 mm for the 24-sided prisms and of 145 mm for the 36-sided prisms, with a thickness of the disks amounting to 22-25 mm.

Absolute measurements of the angles of three 36-sided prisms and one of the two 24-sided prisms (altogether five quartz prisms were produced) were made by means of this equipment. The calibration of these prisms was made by means of two autocollimators by deriving the angles from closed cycle measurements, for which the sum of the prism angles was equal to 360° and the sum of the deviations of these angles from their nominal values was equal to 0°.

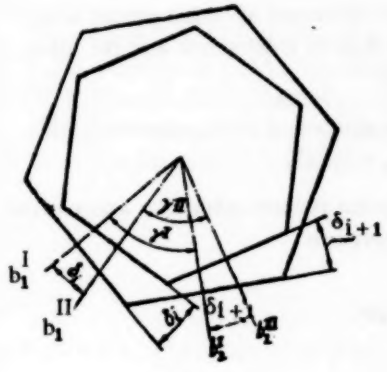

Fig. 1.

The calibration of the prisms and an additional working out of the observed results by means of the dispersional analysis showed that:

1. Systematic errors did not occur during measurements, and random errors followed the normal distribution law.

2. The maximum error of a number of measurements (3σ) approached for any series of measurements the value of 0.5"; hence, for a 36-sided prism the quadratic mean error for 18 series of measurements (of 36 possible series) can be characterized by the value $S = \sigma/\sqrt{18} \simeq 0.04"$.

Each series of measurements consisted of 10 circular (closed) cycles. In passing from one series of measurements to the next the angle between the stationary and rotating autocollimator was changed from 10° (in the first series) to 180° (in the last, 18th series).

3. The technique of making prisms thus developed provided the required accuracy—about half the angles of the prism were made with an error not exceeding 1", and the deviations of the remaining angles are mainly in the limits of 2-3" and do not exceed 4".

Subsequent tests were aimed at determining the accuracy with which the values of the calibrated prism angles can be transmitted to other angle-measuring devices. This problem was tackled by first finding the accuracy of the relative measurements by which the angles of the reference measures were directly compared (collated) with those of the gauge under test. Contrary to absolute measurements in the latter case only one collimator was used.

When two prisms are being compared they are placed one on top of the other on the turntable, and their sides are made to coincide as closely as possible.

The autocollimator is placed in such a manner that the center of its objective is approximately at the same level as the plane separating the two prisms. If the optical axis of the objective is set perpendicularly to the first pair of sides it will be possible to obtain in the instrument's field of vision the images of the autocollimation cross-lines. The azimuth displacement of these two images indicates the displacement of one side of the pair with respect to the other and is measured by means of the eyepiece micrometer, by adjusting in turn to the upper and lower sides. Having rotated the prism through an angle equal to the central angle between the prism sides, similar measurements are made for the next pair of sides, and then these measurements are repeated until the last pair appears before the autocollimator objective.

From these measurements a system of equations can be derived for determining angles δ_i, which indicate the deviation of the prism sides (Fig. 1). If we denote by b_1^I the readings corresponding to the position of the

sides of polyhedron I and by b_i^{II} the readings corresponding to the sides of polyhedron II, we shall obtain

$$\delta_1 = b_1^{II} - b_1^I$$
$$\delta_2 = b_2^{II} - b_2^I$$
$$\cdots\cdots$$
$$\delta_i = b_i^{II} - b_i^I.$$

In these measurements a definite sign convention was adopted. Angles δ_i were considered positive if the autocollimation image from the side of the checked polyhedron was to the right of that obtained from the side of the reference polyhedron, i.e., when $b_i^{II} > b_i^I$. Similarly, for $b_i^{II} < b_i^I$ angles δ_i were considered to be negative. In Fig. 1 the checked polyhedron is shown to be smaller in size than the reference polyhedron and placed above it.

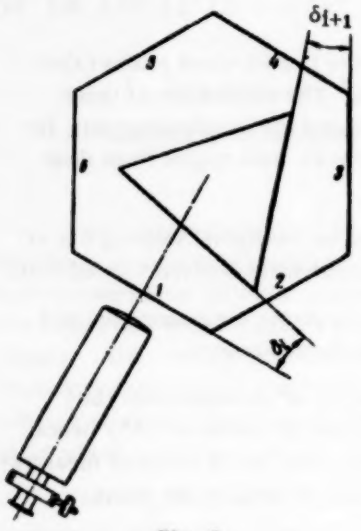

Fig. 2.

The difference $\delta_i - \delta_{i+1} = \Delta_i$ represents the correction to the angle of the checked polyhedron. With this in mind and denoting by γ_i^I the deviations from nominal values of the angles of the reference polyhedron, and by γ_i^{II} the deviations of the angles of the checked polyhedron, we shall obtain for determining γ_i^{II} the equation

$$\gamma_i^{II} = \gamma_i^I + \Delta_i.$$

By this method two previously calibrated 36-sided prisms were compared, one (prism No. 1) being taken as a reference and the other (prism No. 2) as the one being tested.

The quadratic mean error of a number of measurements in this instance was found to be equal to $\sigma_q = 0.25''$.

The certified quadratic mean error (which takes into account the error σ_r of the reference prism) amounted to

$$\sigma_{ct} = \sqrt{\sigma_r^2 + \sigma_q^2} = \sqrt{0.14^2 + 0.25^2} = 0.29''.$$

Finally, this reference equipment was compared with the horizontal interferometer equipment of the VNIIK. For this comparison, which was made for the first time and was of considerable interest, we used a 6-sided steel prism made by the VNIIK and three angle plates of 15, 60 and 75° from the set of VNIIK plates. The angles of the prisms and plates were calibrated by the VNIIK on their interferometer equipment, and then they were measured on the reference autocollimator equipment. In both instances the calibration was made by means of one of the VNIIM 24-sided quartz prisms.

For calibration, the angle plates were placed on the prism in such a manner that their sides were parallel to those of the 24-sided prism wherever their faces corresponded (see Fig. 2 where for simplicity the calibration of a 60° plate by means of a 6-sided prism is shown). Having obtained autocollimation images from the first measured surface of the plate and from the prism side with respect to which this surface was set, readings are taken on the autocollimator eyepiece micrometer (we are considering here the most general case when the measuring surfaces of the plate and the prism side do not completely coincide, i.e., are not in the same plane). Next by turning the polyhedron the second side of the plate is placed in front of the autocollimator and similar readings are taken as before. Next the plate is turned in such a manner that its first measured side comes approximately parallel to the next polyhedron side, which is adjacent to the first. This process is repeated until all the measured surfaces of the plate are made to coincide in turn with all the sides of the prism which is used for calibration.

The comparison of the calibration results of plates with angles of 15, 60 and 75° made by means of the VNIIM reference equipment and the VNIIK interferometer equipment showed that the difference in the value of the angles thus obtained did not exceed 0.3".

The same accuracy was obtained when two angles of a 6-sided steel prism were calibrated against a 24-sided quartz prism. The value of the deviations (γ) of the angles of this prism from their nominal value (60°)

amounted to: as obtained on the reference equipment (VNIIM) $\gamma_1 = -9.5''$, $\gamma_2 = -18.8''$; as obtained on the interferometer equipment (VNIIK) $\gamma_1 = -9.7''$, $\gamma_2 = -18.6''$.

In addition to these measurements the same two angles of the six-sided prism were compared on the reference equipment with ten sides of the 36-sided quartz prism (reference prism No. 1). The following values were then obtained (mean of 10 measurements): $\gamma_1 = -9.9''$, $\gamma_2 = -19.6''$.

The values of γ_1 and γ_2 obtained by comparison of the two methods differ from those obtained on the reference equipment by the calibration method; however, the differences thus obtained are within the tolerances of certification, which take into account the errors of the reference prism calibration and the errors of the comparison method.

The results of the absolute and relative measurement of reference prisms by means of the VNIIM autocollimation equipment, which is intended for the reference method of measuring angles, and the evaluation of the results obtained by comparing this equipment with that of the VNIIK, which is intended (according to the approved test scheme) for certifying 1st grade reference polyhedral prisms, show that the above equipment provides the required accuracy in reproducing and transmitting the value of angles from the highest standard of the reference system.

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IMPROVED TECHNIQUE FOR CHECKING ANGLE-GAUGES BY MEANS OF INSTRUMENT APU-2

S. I. Kuvshinov

Translated from Izmeritel'naya Tekhnika, No. 10, p. 16, October, 1960

In order to raise the accuracy of measurement, decrease the deviations of readings, and simplify the use of the autocollimation instrument APU-2, the author of this article improved the technique of checking with this instrument angle-gauges by means of the comparative method.

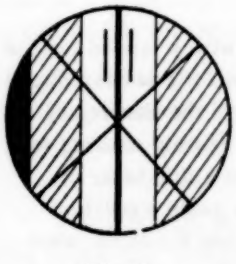

Fig. 1.

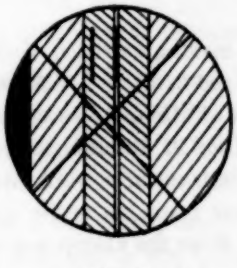

Fig. 2.

When the set is being checked, the lampholder complete with its lamp is lowered to the extreme bottom position of the illuminator body and turned about its axis until in the field of vision there appears (with the angle-gauge in its position) a pale yellow stripe of a width of 1/3 of the field of vision's diameter and a bright yellow central line whose width is roughly equal to that of the sighting hairline, instead of the bright yellow stripe of the same width but with a dark central line, as specified in the instrument description and literature in general. This image is obtained in a field of vision owing to the projection of the hairline side surface instead of the hairline itself.

*See English translation.

This bright yellow, sharp and thin line serves to set, instead of the crossed hairlines (Fig. 1), one of the two vertical lines which are placed above the crossline (we have taken the right-hand line), by the superposition method (Fig. 2).

With this image in the field of vision and such a method of superposition the reading error of the instrument does not exceed 2"; the deviation of readings in ten measurements does not exceed 1" (0.5 of a division). When one and the same angle-gauge is certified by several people the error of measurement lies in the range of 2".

This makes it possible to use instrument APU-2 for certifying 1st-grade angle-gauges with the compulsory condition that the flatness of the angle-gauges should be within 1st-grade tolerances, and the error of the reference angle-gauges used for certifying should not exceed 3".

However, it is advisable to replace the instrument's yellow filter by a pale green one and the three rows of bearing balls by two rows, one for the reference and the other for the measured gauge. The three rows of bearing balls create confusion of the gauges in the sets and lower the accuracy. It is advisable to bear these remarks in view when the instrument is redesigned.

The APU-2 instrument should be supplied to all the State Inspection Laboratories for Measurement Equipment and to the Chelyabinsk Instrument Plant which produces angle-gauges.

When contact instruments are used for checking angle-gauges it should be kept in mind that after checking some 20-25 sets of reference gauges grooves may appear on the working surfaces in the places of contact, thus making further measurements inaccurate. It is, therefore, necessary to check regularly the flatness of the working surfaces of angle-gauges.

MEASUREMENT OF PROFILE ANGLES OF EXTERNAL THREADS BY MEANS OF CASTINGS

M. K. Kovalev

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The profile angle of threads over 100 mm in diameter on nipple locks and ends of tubes has not been checked by the manufacturers up to the present time.

The author of this article has developed a method of checking half the profile angle of external threads over 100 mm. This method consists in reproducing the profile angle of the thread by means of castings with a narrow fin.

The material used for casting consists of medical gypsum in combination with a 2-5% solution of potassium chromate.

Figure 1 shows a threaded tapered plug and a device for making castings. Guide bar 1 with a lapped surface was pressed against the smaller butt end of plug 2 by means of rod 3, strap 4, washer 6 and nut 5. The plug thread was covered by mold 6, which was fixed to the guide by means of clamp 7. In order to prevent the skewing of surface A, a piece of rubber tubing 8 was slipped on the end of rod 3. In order to prevent the casting sticking to the thread of the gauge, the walls of the mold and the guide, their surfaces were smeared with a thin layer of transformer oil. As soon as the casting became hard, the rod was taken out of the guide, the guide was lifted and simultaneously withdrawn from the butt end of the gauge. Next the clamp was loosened and the mold taken off the guide. The mold was then gently forced open by hand and the casting deposited on a piece of paper or a clean cloth, the burrs were cleaned off the edges of surface B (Fig. 2) and the casting placed in a holder (Fig. 3). Surface B formed by the casting on plane A of guide 1 is used as the base of the casting (Fig. 1). By tightening screw 1 (Fig. 3) casting 2 is pressed against the opposite plane without allowing it to skew. The device is placed

between centers 3 and 4 of a universal microscope and half the angle is determined from the large and small sides of the profile of two roots (a and b) in the middle of the casting.

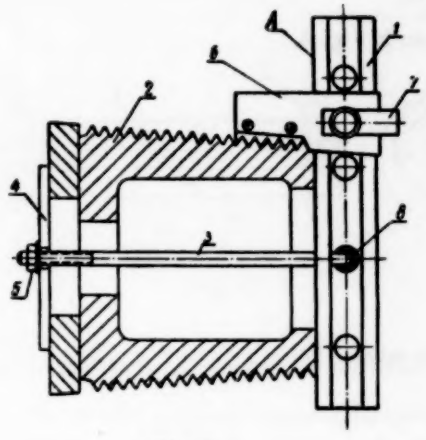

Fig. 1.

Fig. 2.

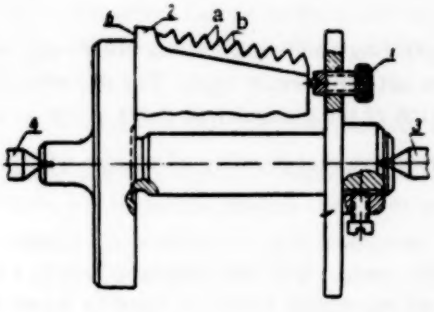

Fig. 3.

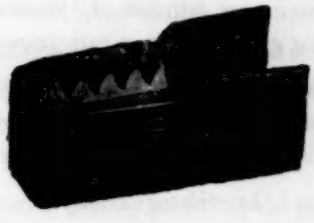

Fig. 4.

The accuracy of reproducing the profile angle of an outside thread by means of a casting was checked. The profile angle of the gauge was measured by a special Krysanov microscope equipped with measuring blades. The profile angle of two roots in the casting was determined by measuring three times each half-angle without the help of blades. The mean of the three measurements of two roots was taken as the measurement result. For this purpose the profile angles of the gauge thread were compared with the corresponding angles of the casting.

The error in reproducing the profile angle by means of castings made of gypsum with 2-5% of potassium bichromate amounted to −10' to +15' for the full profile angle of threads with tolerances for half a profile angle, of ±45' in the case of locks and ±75' in the case of boring or drive pipes.

The accuracy in reproducing the profile angle also depends on the thickness of the casting rib C (Fig. 2). The thicker the rib the less accurate becomes the profile angle. On the other hand, a very thin rib breaks easily when it is being taken out of the mold, and the gypsum does not completely fill the roots of the thread. We established by means of tests the optimum rib thickness for pipes and locks of various diameters:

for diameters over 100 and up to 180 mm it is 2.75 mm;
for diameters between 180 and 245 mm it is 3.25 mm;
for diameters between 273 and 325 mm it is 3.75 mm;
for diameters over 325 and up to 500 mm it is 4.25 mm.

Plane D (Fig. 2), along which the profile angle is measured, is displaced from the center by half the thickness of the rib. The systematic error due to this displacement does not exceed -1'30" in half the profile angle of a thread of a $8^{9}/_{16}$" plug gauge (its diameter is 127 mm at the small butt end). For larger diameters this error decreases, hence it can be ignored.

The preparation of the mixture of gypsum with a solution of potassium bichromate, the filling of the mold by the mixture and the hardening of the cast take about 5-6 minutes.

The mixture proportions for gypsum casts are: medical gypsum with a 2-5% solution of potassium bichromate 1: 1; medical gypsum with a 5% solution of sodium nitrite 6.5: 3.5.

The construction of the mold with the casting placed in it is shown in Fig. 4.

CHECKING SMALL DIAMETERS

A. V. Vysotskii

Translated from Izmeritel'naya Tekhnika, No. 10, pp. 18-19, October, 1960

Checking finished axles of small diameters (2-4 mm) is made by means of normal measuring equipment and does not present any difficulties. However, the use of ordinary continuous inspection instruments for measuring these axles during grinding is as a rule impossible.

Caliper gauges are too heavy for continuous checking during grinding and require a relatively large measuring effort, thus leading to the bending of small diameter axles which are not sufficiently rigid. The deformation of the axles interferes with the grinding process and leads to the distortion of the geometrical shape of the axles.

The Bureau of Interchangeability constructed for checking during external grinding of articles with thin mandrels (from 1 mm) a special instrument type BV-4003.

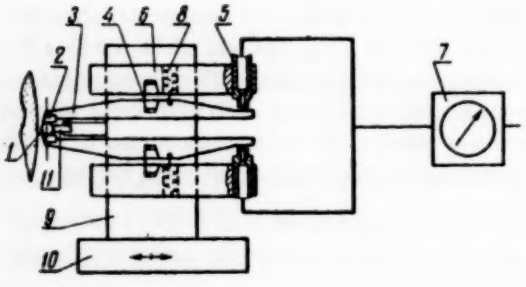

Fig. 1.

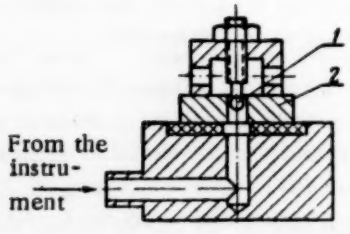

Fig. 2.

The instrument (Fig. 1) consists of a caliper gauge which makes contact with the measured article 1 by means of two measuring feelers 2, fixed to levers 3 with independent suspension on flat springs 4. At their other end the levers have flat surfaces placed opposite the butts of two measuring nozzles 5. These nozzles are mounted on carriages 6, on which the levers are suspended. The total gap of the two nozzles, and hence the readings of the pneumatic instrument 7, depend on the size of the measured axle.

The use of two independently suspended light measuring levers and the pneumatic method of registration make it possible to lower considerably the measuring effort provided by springs 8, thus avoiding the bending of the machined details.

The readjustment for various sizes is accomplished by the displacement of carriage 6 along stand 9. The stand is fixed to base 10, which serves for approaching and withdrawing the caliper gauge.

Stop 11, which serves as a guide, can be fixed to the stand. For checking small diameter holes (1-4 mm) under production conditions, the Bureau of Interchangeability developed and investigated two methods based on pneumatic measurements which can easily be made automatic.

The first method uses direct transmission of air through the measured hole. In order to obtain a sufficiently large sensitivity, a calibrated ball (Fig. 2) is placed in the hole and the circular gap formed between ball 1 and hole 2 is then measured.

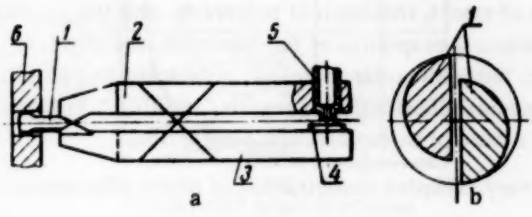

Fig. 3.

Tests carried out on a 3.2 mm hole showed that with a correctly chosen ratio of the nozzle diameter to that of the ball it is possible to obtain a high transmission ratio (in a high pressure instrument with a water manometer it amounted up to 30,000). The reading is stable providing the ball touches during measurement the wall of the hole, thus providing a constant shape for the gap.

However, this method cannot be used for checking blind holes and determining the deviation of the hole from the correct geometrical shape. Another shortcoming of this method consists in the difficulty of obtaining a hermetic seal between the butt of the detail and the device through which the air is supplied.

The second method, a pneumatic-contact method, is free from these defects. For experimental purposes we produced a device (Fig. 3a), whose measuring feelers 1, used for checking 1 mm diameter holes, were fixed to levers 2 and 3, which could freely rotate on a cross-shaped joint made of flat springs.

One of the levers carries buffer 4 and the other, nozzle 5. The nozzle gap depends on the dimension of hole 6 in the detail. In order to obtain maximum rigidity the measuring probes 1 are placed in the manner shown in Fig. 3b.

Tests have shown that the measurement error of this instrument amounts to 0.0005 mm.

The defects of this method include the necessity of having additional setting facilities for the measured detail, since the measuring feelers cannot be used for centering owing to their insufficient rigidity.

INVESTIGATION OF THE RINGING CAPACITY OF BLOCK GAUGES

Lu Ch'ao-tsêng and A. S. Akhmatov

Translated from Izmeritel'naya Tekhnika, No. 10, pp. 19-22, October, 1960

The ringing capacity of block gauges (the capacity to stick together and form blocks) is important from the point of view of a firm connection, ease with which it can be attained and stability in use, but it is not a decisive technical problem.

The sticking together of plates* in lapping can be evaluated by their resistance to tangential displacement (by static friction), in the manner specified by GOST 9038-59 (State Standard), or by the force (work) required to separate the plates (providing the elementary forces acting normally to the plane of interaction are uniformly distributed and their resultant is exactly in the center of the plate) [1].

In [2] both categories of forces were measured simultaneously with respect to several variables in order to discover the most important of them which have a primary technical or physical significance. The ringing

*Henceforth we shall use the terminology employed in physical literature: "plates" (block gauges), "static friction" (capacity to ring), "adhesion of plates" (ringing, sticking together), etc.

capacity of a plate depends on three basic factors: 1) the properties of the metal and its surface; 2) the properties of the lubricating mixture and its molecules; 3) the properties of the intermediate layer in the kinetic process of its formation and under balanced, stationary conditions. The functional relations of the ringing capacity in each of these categories are very numerous and have distinctly unequal values. Let us enumerate those which the authors of this article aimed at investigating: 1) the effect of the microgeometrical profile, the metal structure and its defects, hardness, the physical nature of the metal (the grade of steel), mechanical properties, and the physicochemical purity of the surface; 2) the volumetric physicochemical properties of the lubricant, the physical properties of its molecules, their atomic structure, isomerism, intramolecular mobility, polarizability, polarity, and chemical activity; 3) the method of obtaining an intermediate lubrication layer, its "training," "thickness," physical condition, structure, mechanical properties, and the nature of intermolecular interactions.

Thus, the ringing ability of block gauges depends on a very complex combination of physical phenomena and processes.

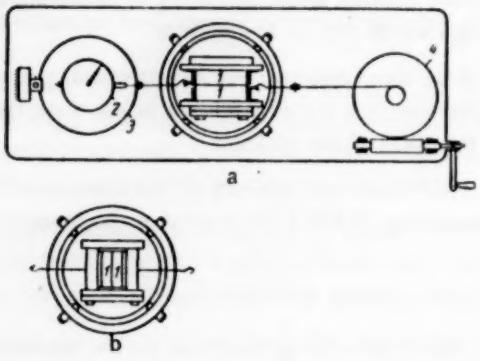

Fig. 1. Schematic of a friction-adhesion gauge. a) Connection for measuring friction; b) connection for measuring adhesion; 1) plates under investigation; 2) indicator; 3) dynamometer; 4) extending mechanism.

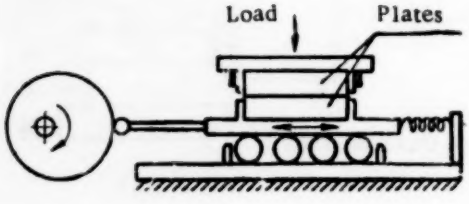

Fig. 2.

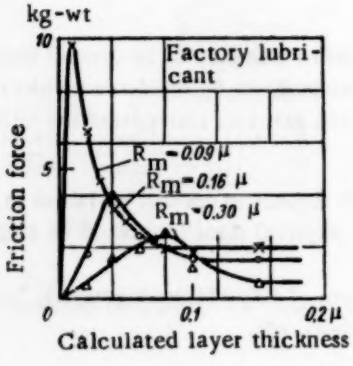

Fig. 3.

Technique. The greater part of the friction and adhesion measurements were made by means of a friction-adhesion gauge of our own construction (Fig. 1), which is a universal, handy, reliable and sufficiently accurate instrument.

In order to account for the wear of plates in lapping, a special device was constructed (Fig. 2) with the main object of a sufficiently full simulation of plate lapping under conditions of normal use (hand lapping).

The surfaces were cleaned either with pure benzine or by an adsorption-abrasive powder (coal, silica gel). Friction coefficient measurements served as a criterion of the surface cleanness.

The "Kalibr" Plant plates were studied and compared with those made by Krasnyi Instrumental'shchik, Harbin, Ford, Zeiss, Johansson and others.

Static friction, adhesion and wear without lubrication. For physicochemically extremely pure plates friction (without a load) and adhesion are very small, whereas wear is catastrophically large.

Static friction of plates with lubrication. Curves shown in Figs. 3, 4 and 5 illustrate the test results. All the curves (at high surface smoothness) have a maximum, whose position does not depend on the nature of the lubricant, but is displaced for any given lubricant (Fig. 5) and a rising surface roughness toward greater layer thickness. The absolute size of the maxima varies; its largest value was observed for a factory lubricant and a surface finish grade 13, at a layer thickness of 0.01 μ.

Adhesion of lubricated plates. Figure 6 shows test results obtained for the value of adhesion with respect to layer thickness for various grades of surface roughness. These curves remained of the same shape for all the investigated lubricants. The maximum adhesion (similarly to the friction force) was observed for a factory lubricant and the maximum grade of smoothness approximately at the same layer thickness of 0.01 μ. The point of inflection of the curves (irrespective of the type of lubrication approximately constant for each given grade of smoothness)

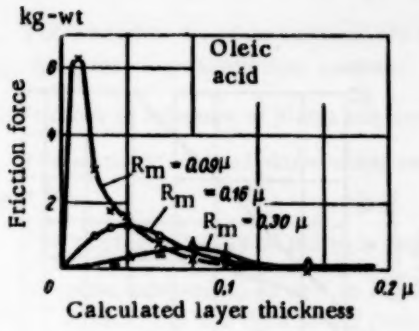

Fig. 4.

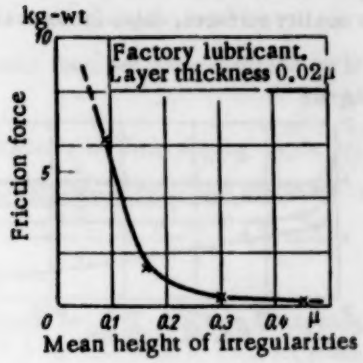

Fig. 5.

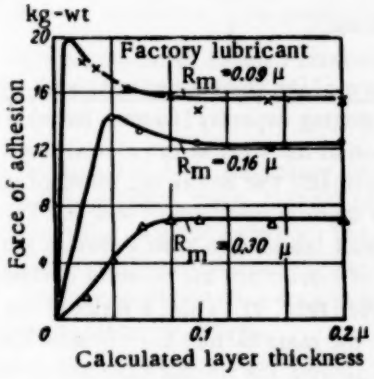

Fig. 6.

is displaced with a decreasing degree of smoothness toward the larger layer thicknesses.

Analytical form of the relation of the boundary friction forces in the plate to loading (application of B. V. Deryagin's binomial formula). We studied the relation of the friction forces to loading (up to 40 kg-wt) in plates (grade 13b) with various lubricants. The measurement results can be expressed by the binomial friction formula $f = \mu (sp_0 + N)$ suggested by B. V. Deryagin. The value of sp_0, which represents the adhesion of the plates, was found by means of extrapolating these curves. It was found to equal the value of adhesion obtained by direct measurement within the limits of measurement errors (see table).

Lubrication	Force of sticking (adhesion), kg-wt		Actual friction coefficient μ	Specific pressure of molecular adhesion p_0, kg-wt/cm²
	extrapolation result	direct measurements		
Oleic acid	15.5	16.0	0.40	5.7
Factory lubricant	22.0	20.0	0.45	8.2
Vaseline oil	19.0	18.0	0.45	7.0

P. V. Denisov [3] arrived at similar conclusions on the basis of different conditions of observation. Thus, there remains little doubt that under the conditions of our experiments the first term of the above formula is due to adhesion.

Since under our conditions the nominal contact area is equal to the actual one, we were also able to find such interesting quantities as the "real" friction coefficient (μ) and the specific molecular adhesion (p_0).

The effect of relatively high loading on the boundary friction forces in the plate and on the structure of the lubricating layer. The decreasing friction coefficient at pressures above 8 kg-wt/cm² is probably due to the transformation of the lubricating layer structure, and to the variation (increase) under these conditions of the layer's elastic constants, a variation discovered by L. V. Panova and A. S. Akhmatov by means of the "pile of layers" method.

Effect of hardness and the grade of steel. The friction force of plates lapped with a factory lubricant rises proportionately to their hardness. The effect of the above factors on the friction and adhesion of lubricated plates

is probably indirect-in the same manner as the chemical composition of steel, its structure and hardness help to obtain high quality surfaces, these factors raise the ringing capacity of plates.

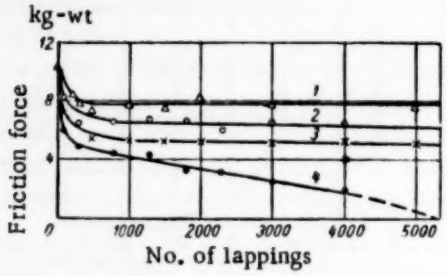

Fig. 7. 1) Oleic acid; 2) factory lubricant; 3) vaseline oil; 4) dry lapping.

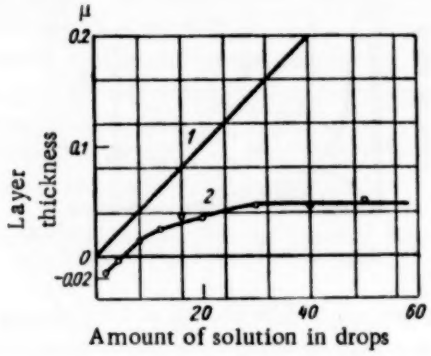

Fig. 8. 1) Calculated thickness; 2) actual thickness.

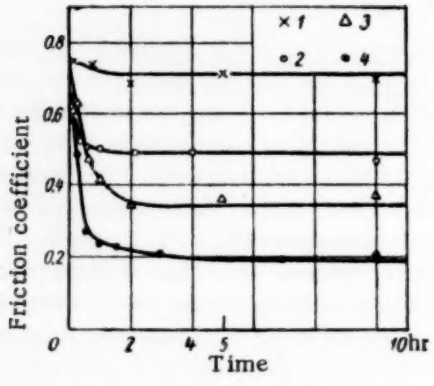

Fig. 9. 1) Reference plate; 2) vaseline oil; 3) oleic acid; 4) stearic acid.

Variation of the ringing capacity with wear. The value of the ringing capacity (friction force) was taken as the criterion of the wear in the plates, since this capacity is normally lost before the changed size of the plates affects them as standards of length. The curves of the wear without lubrication have a normal shape; in the presence of lubrication the phase of catastrophic wear was not observed (Fig. 7). Thus, a rational use of lubrication is the main factor, as compared with all the others, for preserving the ringing capacity of the plates in use.

The effect of the lubrication layer thickness on the dimension of the gauge block (measurements of the actual lubrication layer thickness). The size of 10 plates was measured without lubrication, and the size of the block formed by them with lubrication and good ringing conditions. The actual thickness of a single lubrication layer between the plates used as elements of a single block does not affect substantially the size of the block (Fig. 8). The effect of "negative" thickness layers is due to the removal of the microgeometric profiles of the plates.

Effect of absorption and liberation of the lubricants by the plates. In view of the modern conception of the existence inside metals of a microcapillary network of cavities and cracks, and the observations made by N. P. Petrov, we attempted to introduce the lubricant inside the plates of the "Kalibr" Plant and to observe its spontaneous liberation. It is possible to introduce lubricants inside a steel block, for instance by means of periodic raising and lowering of temperatures. For this purpose the plates, placed in baths of various lubricants, were periodically heated up to 90-95°C and then cooled; this heat treatment lasted 20 days. These plates were then carefully cleaned and their surfaces tested for friction (the variation of the friction coefficient with time was determined). The results of these experiments are shown in Fig. 9. These experiments revealed a spontaneous liberation of lubricants from the body of the plate to its surface. The amount of liberated lubricant was, however, insufficient for a satisfactory adhesion of the plates.

SUMMARY

1. A thin lubricating layer on the surface of block gauges is a prerequisite for their effective ringing capacity.

2. The thickness of this layer, under standard conditions of ringing, which provides a maximum adhesion of plates with a high grade of smoothness (13b), is constant, and equals approximately 0.01 μ. The variations in the

length of a gauge block due to the lubricating layers is a multiple of their number and thus will lie within tolerances.

3. The smoother the plate surfaces are made, the firmer becomes their adhesion obtained by lapping, providing the remaining factors stay constant.

4. Friction or adhesion of plates can serve equally well as a criterion of their ringing capacity.

5. The static friction of plates under conditions of boundary lubrication follows the binomial law $f = \mu(sp_0 + N)$.

6. The ringing capacity of plates is proportional to their hardness.

7. Nonpolar substances, as well as highly polar organic substances, are less effective in lapping than lubricating compounds of the type of the factory lubricant "Kalibr." It is advisable to conduct research for new lubricants by mixing polar and nonpolar substances.

8. The firmness of adhesion between the plates is determined by the physicochemical properties of chain molecules in the lubricants, the energy of their adhesive and cohesive interaction, the peculiarities of the layer structure, and the variations of its properties as a function of the distance between the solid surfaces.

9. It is expedient to conduct research into the possibility of self-lubrication of the plates.

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THE THEORY OF AUTOMATIC BEAM SCALES WITH ELECTRICAL BALANCING

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At present there is a growing demand in scales for measuring rapidly changing weights and other rapidly variable forces. Ordinary beam scales with balancing by means of weights are unsuitable for this purpose. For such purposes beam scales with an electrical meter and balancer are often used. The use of an electrical balancer changes all the characteristics of the scales, such as the beam balancing equation, damping conditions and sensitivity. The sources of weighing errors in scales with an electrical balancer will differ from those of ordinary scales.

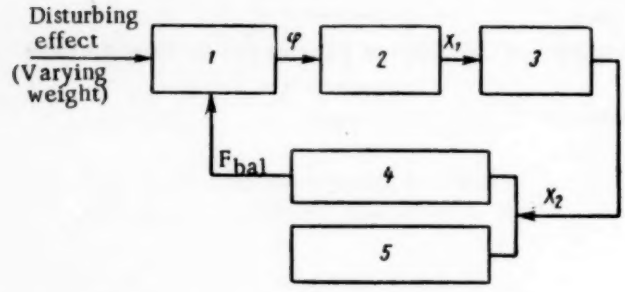

Fig. 1. 1) Mechanical scales (object of control); 2) sensing element; 3) amplifier; 4) controlling element (weight balancer); 5) weight measuring instrument.

The present article deals with certain principles of the theory of beam scales with electrical balancing.

The task of the electrical device added to the beam scales for measuring variable weights consists in bringing the scale beams to a balanced position, i.e., the electrical device serves as an automatic regulator of the balanced position of the scales. The measured weight can, in this case, be considered as a disturbing effect, which unbalances the controlled object, the scales. The functions of the automatic regulator of the scales are extended by the necessity of measuring the disturbing effect, the varying weight.

In order to derive the equation of the beam movement and the formula of sensitivity, and investigate the conditions of damping of scales with electrical balancing, it is, therefore, possible to apply the theory of automatic control systems.

The block schematic of any scales with electrical balancing can be represented in the form shown in Fig. 1.

The controlled value in the scales consists of the angle of deviation of the beam from its balanced position. The electrical regulator incorporated in the scales must bring them to balance. As sensing elements reacting to the imbalance of the automatic scales may be used any electrical circuits with displacement transducers, such as ionization, photoelectric, capacitative, inductive and other transducers [1-4]. The basic requirements for such transducers consist of a linear characteristic and a minimum reaction on the scale beam.

The output variable of a sensing element is represented by some electrical quantity X_1, which depends on the type of the displacement transducer used.

Since the signal from the sensing element is very weak, it is necessary to place between it and the controlling device, which actuates the scale beam, an amplifier whose type depends on the transducer used.

Electromagnetic balancing mechanisms, consisting of a permanent magnet suspended from the scale beam and an electromagnet, are more frequently used than any other regulating or balancing devices.

The peculiarities of an automatic control system for scales consist of the following: In the majority of cases the inertia of the electrical regulator incorporated in the scales is negligible as compared with that of the controlled object, consisting of the mechanical scales, and the regulator can be considered as an ideal element [5]. For convenience of recording the variable weight, the automatic control system of the scales must have an aperiodic transient process. In order to obtain an aperiodic transient process it is necessary in certain cases to introduce a derivative into the law of regulation; then the regulator will become an ideal element with the introduction of a derivative [5].

In deriving dynamic equations for the automatic control system of the scales it is, therefore, sufficient to examine an ideal regulator and an ideal regulator with the introduction of a derivative.

From the point of view of the nature of the disturbing effect it is necessary to examine two cases:

1) disturbing effect changes instantaneously from 0 to a certain determined value;

2) the disturbing effect increases smoothly with time at a constant rate of $P = V_0 t$.

Let us now derive a dynamic equation for an automatic control system of the scales, examine the conditions of the system's stability, and the damping conditions.

The dynamic equation of the controlled object (beam scales) will be the same for any beam scales. The input variables of this automatic control system element consist of the measured weight (disturbing effect) P and the balancing force F_{bal}, and the output variable will be the beam's angle of deflection φ. The dynamic equation of this element must relate quantities P, φ and F_{bal} with the scale parameters.

In order to derive the dynamic equation of the controlled object let us use the Lagrange equation [6].

Let us take the beam deviation angle as a generalized coordinate, and the derivative of this angle with respect to time as a generalized velocity. By computations similar to those used in [6] we obtain *

$$J_{rf}\ddot{\varphi}+\left(\frac{P'}{2g}l^2+\mu_{rf}\right)\dot{\varphi}+P_1C\varphi=-F_{bal}\,l+Pl, \tag{1}$$

where J_{rf} is the referred moment of inertia of the scales; μ_{rf} is the referred coefficient of viscous friction; P' is the speed of variation of the measured weight; P_1 is the weight of the balance; C is the distance between the beam's center of gravity and the fulcrum line; l is the length of the beam arm; F_{bal} is the electrical balancing force; P is the measured weight.

Having divided all the terms of (1) by P_1C and denoted the time constant and the gain of the controlled object by

$$\frac{J_{rf}}{P_1C}=T_2^2;\qquad \frac{1}{P_1C}\left(\frac{P'}{2g}l^2+\mu_{rf}\right)=T_1;\ \frac{l}{P_1C}=K, \tag{2}$$

we can convert (1) to a standard form

$$T_2^2\ddot{\varphi}+T_1\dot{\varphi}+\varphi=-KF_{bal}+KP. \tag{3}$$

Thus (3) is the dynamic equation of the controlled object.

The static characteristic of the controlled object (scales) is expressed by the formula

$$\varphi=KF_{bal}, \tag{4}$$

*In further considerations this equation is assumed to be linear, since the variation of the coefficients of the derivatives is very small. For instance, the variation of the system's moment of inertia due to the variation of its mass amounts to about 1% for 200-gram scales.

where K is determined from (2) and represents the slope of the static characteristic of the controlled object, or the gain of the object.

It follows from (2) that the value of K represents the mechanical sensitivity of the scales and depends only on the parameters of the beam l, P_1 and C.

Let us now examine the dynamic characteristics of the element in a transient state (self-oscillations of the element). Let us also assume that the disturbing effort does not exist, i.e., that P = 0. Equation (3) will then have the form

$$T_2^2\ddot{\varphi}+T_1\dot{\varphi}+\varphi=-KF_{bal}. \tag{5}$$

The controlled object is an aperiodic element of the second order for $T_1 \geq 2T_2^2$ and an oscillating element if $T_1 < 2T_2^2$. Time constant T_1 represents the damping of the element's self-oscillations, and time constant T_2^2 represents their building up.

It follows from (2) that time constant T_1 depends on the speed of variation of the measured weight, and increases with the rise of the latter. Hence, the measured weight itself serves to damp the scales. For scales of a small weighing capacity, which have a small referred moment of inertia J_{rf}, this type of damping may not be sufficient in order to obtain an aperiodic transient process.

Time constants T_1 and T_2^2 and the gain K of the object can be easily calculated from (2).

In certain cases it is easier to determine the value of T_1 graphically instead of calculating it from (2). For this purpose it is necessary to plot the transient characteristic of the scales as the object of control.

Investigation of the automatic control system of the scales. It is sufficient to examine the incorporation in the scales of two types of regulators, an ideal regulator whose dynamic equation is

$$F_{bal}=K_{rg}\varphi, \tag{6}$$

and an ideal regulator with the introduction of a derivative, whose dynamic equation is

$$F_{bal}=K_{rg}\varphi+K_{rg}K_0\dot{\varphi}, \tag{7}$$

where K_{rg} is the regulator gain equal to the product of the gains of all the regulator elements-the sensing element, the amplifier and the control element (Fig. 1), and K_0 is the coefficient representing the intensity of introducing a derivative into the control law. In case when an ideal regulator with the introduction of a derivative is incorporated in the scales, the dynamic equation of the automatic control system assumes the form

$$T_2^2\ddot{\varphi}+(T_1+KK_0K_{rg})\dot{\varphi}+(1+KK_{rg})\varphi=KP. \tag{8}$$

The system is stable if the coefficients in the left-hand side of (8) are positive.

The control process consists of the transient and the stable state periods:

$$\Delta\varphi=\Delta\varphi_{tr}+\Delta\varphi_{st}. \tag{9}$$

The transient process is determined by the general solution of the homogeneous Eq. (8) for P = 0, and the stable state process by a particular solution of the nonhomogeneous Eq. (8) for a given right-hand side part of P. In order to determine the conditions of damping and calculate the settling time of the beam oscillations let us examine the transient processes.

It can be easily shown that for $4T_2^2(1+KK_{rg}) > (T_1+KK_0K_{rg})^2$ the transient process is oscillatory and is determined from the formula

$$\Delta\varphi_{tr}=C_1e^{-\frac{t}{T_a}}\sin(\omega t+C_2), \tag{10}$$

where

$$T_a = \frac{2T_2^2}{T_1 + KK_0K_{rg}};$$

$$\omega = \sqrt{\frac{1+KK_{rg}}{T_2^2} - \left(\frac{T_1+KK_0K_{rg}}{2T_2^2}\right)^2},$$

and the arbitrary constants C_1 and C_2 can be determined from the initial conditions (for $t = 0, \Delta\varphi_{tr} = C$ and $\varphi = D$) and are equal to

$$C_1 = C\sqrt{1+\frac{1}{\omega^2}\left(\frac{D}{C}+\frac{1}{T_a}\right)^2};$$

$$C_2 = \sin^{-1}\frac{C}{C_1}.$$

The duration of the transient process is

$$t_c \approx 3T_a.$$

The controlled object without the regulator has a time constant of the exponent $T = 2T_2^2/T_1$. Hence, the incorporation of a regulator in the scales decreases the time constant of the exponent to an increasing degree with a rising regulator gain and a rising intensity in introducing a derivative in the law of control.

With a rising K_{rg} a moment will arise when the following equality will hold:

$$4T_2^2(1+KK_{rg}) = (T_1+KK_0K_{rg})^2.$$

The transient process will then be aperiodic:

$$\Delta\varphi_{tr} = (C_1+C_2t)e^{-\frac{t}{T_a}}, \tag{11}$$

where $T_a = (2T_2^2)/(T_1 + KK_0K_{rg})$ and the arbitrary constants C_1 and C_2, which can be obtained from the initial conditions, are equal to $C_1 = C$ and $C_2 = D + C/T_a$.

With a further rise in K_{rg} the following inequality will hold:

$$4T_2^2(1+KK_{rg}) < (T_1+KK_0K_{rg})^2.$$

The transient aperiodic process

$$\Delta\varphi_{tr} = C_1e^{-\frac{t}{T_a}} + C_2e^{-\frac{t}{T_b}} \tag{12}$$

consists of two exponents with different time constants:

$$T_a = \frac{2T_2^2}{T_1+KK_0K_{rg} - \sqrt{(T_1+KK_0K_{rg})^2 - 4T_2^2(1+KK_{rg})}};$$

$$T_b = \frac{2T_2^2}{T_1+KK_0K_{rg} + \sqrt{(T_1+KK_0K_{rg})^2 - 4T_2^2(1+KK_{rg})}}.$$

The arbitrary constants C_1 and C_2 are determined from the initial conditions and equal

$$C_1 = \frac{T_a}{T_a - T_b}(C - T_b D);$$

$$C_2 = -\frac{T_b}{T_a - T_b}(C + T_a D).$$

Above we have investigated the relation between the transient process in the control system and the regulator gain K_{rg}. The scales regulator should be constructed in such a manner that the transient process is aperiodic, and the duration of the transient process, determined by time constants T_a and T_b, should be as short as possible. Figure 2 represents the relation of the time constants T_a and T_b to the regulator gain.

To obtain a small settling time of the scale beam oscillations it is advisable to work with gain K_{rg} lying within the aperiodic range. However, this value of the gain should be coordinated with the scale's mechanical sensitivity which, as it will be shown, is determined by the gain of the electrical regulator.

In order to determine the deviation angle of the beam at the time of balancing, let us examine the stable state process, determined by the particular solution of the nonhomogeneous Eq. (8) for a given right-hand side of $P = P_0$ and $P = V_0 t_c$. For the first case the particular solution of (8) takes the form of a static deviation

$$\varphi_{st} = \frac{KP_0}{1 + KK_{rg}} \tag{13}$$

and for the second case

$$\varphi_{st} = \frac{KV_0 t_c}{1 + KK_{rg}}. \tag{14}$$

The static angle is proportional to the measured weight, and the coefficient of proportionality between P and φ_{st},

$$E = \frac{K}{1 + KK_{rg}}, \tag{15}$$

represents the mechanical sensitivity of the scales. Equations (13) and (14) can be rewritten in the form

$$P_0 = F_{bal} + \frac{\varphi_{st} P_1 C}{l}; \tag{16}$$

$$V_0 t_c = F_{bal} + \frac{\varphi_{st} P_1 C}{l},$$

where

$$F_{bal} = K_{rg}\varphi_{st},$$

$$K = \frac{l}{P_1 C}.$$

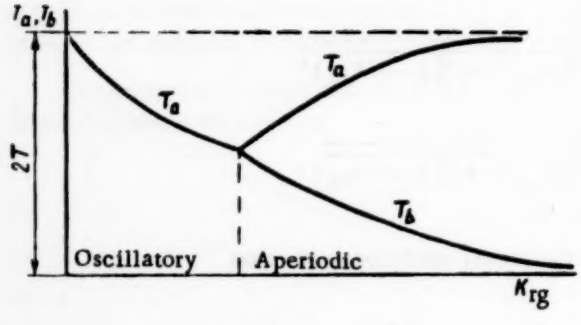

Fig. 2.

Hence the electrical balancing force F_{bal} does not completely balance the measured weight at the instant the scales are balanced and the weight recorded. The weight is partly balanced by the moment which arises because the center of gravity of the scale beam does not completely coincide with the fulcrum line of the beam. The value of the electrical out-of-balance is expressed by the second term of (16), depends on the weight of the beam P_1, and, hence, will differ in different parts of the earth's surface. The weight is registered on scales with electrical balancing by means of an electrical measuring instrument, which measures or records the

electrical quantity proportional to the electrical balancing force. A calibration characteristic is obtained which expresses the relation between the electrical quantity and the mass of reference weights, which are placed on the pan of the scale during calibration. The electrical out-of-balance is then automatically accounted for. However, the calibration characteristic will only hold for that place on the earth's surface in which the scales were calibrated. Elsewhere the weight P_1 of the beam will be different owing to the variation of the gravitational force; hence, the value of the electrical out-of-balance will also be different. In order to eliminate this out-of-balance it is sufficient to adjust the scale beam in such a manner that its center of gravity coincides with the fulcrum line (C = 0).

By inserting in (16) C = 0 we obtain

$$\begin{aligned} P_0 &= F_{\text{bal}}\ ; \\ V_0 i_c &= F_{\text{bal}}\ . \end{aligned} \tag{17}$$

Thus for C = 0 the electrical force completely balances out the weight of the point of equilibrium. In this instance the scales will be in an unstable position without the regulator, since the center of gravity of the vibrating system coincides with the fulcrum line.

The dynamic equation of the controlled object has the form

$$T\ddot{\varphi} + \dot{\varphi} = -K_1 F_{\text{bal}} + K_1 P\ , \tag{18}$$

where

$$T = \frac{J_{\text{tr}}}{\frac{P'}{2g} l + \mu_{\text{tr}}}\ ;$$

$$K_1 = \frac{l}{\frac{P'}{2g} l^2 + \mu_{\text{tr}}}\ .$$

Let us examine the case when a regulator with an introduced derivative is incorporated in the scales. The dynamic equation of the automatic control system is then determined by

$$T\ddot{\varphi} + (1 + K_1 K_0 K_{\text{rg}})\dot{\varphi} + K_1 K_{\text{rg}}\varphi = K_1 P. \tag{19}$$

The operation of the automatic control system, as in the previous case, is represented by a second order differential equation. Equations (19) and (8) only differ by the coefficients of variable φ and its derivatives.

Since the coefficients of φ'', φ', and φ are positive, the system is stable, i.e., the incorporation of an electrical regulator in the scales with an unstable balanced position (to a "neutral" object of control) makes it stable. Moreover, by selecting the regulator gain it is possible to make the scales operate as an oscillating or an aperiodic control system.

In order to determine the conditions of damping and calculate the settling time of the beam oscillations let us examine the transient processes.

It can be easily shown that for $4TK_1K_{\text{rg}} > (1 + K_0K_1K_{\text{rg}})^2$ the transient oscillatory process is determined by (10), where

$$T_a = \frac{2T}{1 + K_0 K_1 K_{\text{rg}}}\ ;\quad \omega = \sqrt{\frac{K_1 K_{\text{rg}}}{T} - \left(\frac{1 + K_0 K_1 K_{\text{rg}}}{2T}\right)^2}\ .$$

For $4TK_1K_{\text{rg}} = (1 + K_0K_1K_{\text{rg}})^2$ the boundary of aperiodicity is reached. The transient process in this instance is determined by (11), where $T_a = (2T)/(1 + K_0K_1K_{\text{rg}})$.

For $4TK_1K_{\text{rg}} < (1 + K_0K_1K_{\text{rg}})^2$ the aperiodic transient process is determined by (12), where

$$T_a = \frac{2T}{(1+K_0K_1K_{rg}) - \sqrt{(1+K_0K_1K_{rg})^2 - 4TK_1K_{rg}}};$$

$$T_b = \frac{2T}{(1+K_0K_1K_{rg}) + \sqrt{(1+K_0K_1K_{rg})^2 - 4TK_1K_{rg}}}.$$

The deviation angle of the scale beam is determined at the instant of balance by the particular solution of the nonhomogeneous Eq. (19) for a given right-hand part of $P = P_0$ and $P = V_0 t_c$.

For the first case a particular solution of (19) will take the form of a static deviation $\varphi_{st} = P/K_{rg}$ and for the second case $\varphi_{st} = V_0 t_c / K_{rg}$.

Investigating the formula for the mechanical sensitivity of scales. The mechanical sensitivity K of scales without an electrical regulator of the balance position is determined from (2) and depends on the scale beam parameters l, P_1 and C.

The mechanical sensitivity E of scales with an electrical regulator of the balance position is determined from (15). Inserting in (15) $K = l/P_1C$, we obtain

$$E = \frac{l}{P_1C + lK_{rg}}. \tag{20}$$

It follows from (20) that the mechanical sensitivity of scales with an electrical regulator depends not only on the beam parameters, but also on those of the electrical regulator, and that with a rising K_{rg} the sensitivity decreases.

It has previously been found that scales supplied with an electrical regulator of the balance position should be adjusted in such a manner that the scale beam center of gravity coincides with the fulcrum line (C = 0).

Inserting in (20) C = 0 we obtain

$$E_1 = \frac{1}{K_{rg}}. \tag{21}$$

It follows from (21) that for C = 0 the mechanical sensitivity of the scales does not depend on the beam parameters and is completely determined by the parameters of the electrical regulator K_{rg}.

SUMMARY

By means of the formulas derived in this article it is possible to determine the movement of the beam in a transient process, calculate the settling time of the beam oscillations and the angle of deviation of the beam at the instant of balancing, as well as the mechanical sensitivity of the scales for given beam parameters and those of the electrical regulator of the balance position.

Above formulas can be used for designing scales with electrical regulators of the balance position.

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MECHANISM FOR MOVING A REFERENCE-WEIGHT FLATCAR

V. I. Malyshev

Translated from Izmeritel'naya Tekhnika, No. 10, p. 27, October, 1960

For adjusting and checking railroad car weighbridges of a 100-ton capacity, 25-ton reference-weight flatcars are normally used.

During testing the flatcar is pushed onto the bridge in a certain direction and brought to rest in eight different positions; then the direction is reversed and the same process repeated eight times. The flatcar is pushed onto the bridge several times.

This labor-consuming process was until recently carried out manually by 3-4 men. The flatcar was positioned on top of the bridge supports by means of parking brake blocks, which stopped it with a jerk, thus distorting the weighbridge reading and making its adjustment more difficult.

In order to make this operation easier and raise the productivity of labor, the author of this article has developed, for the purpose of moving the flatcar, a simple mechanism which consists of a reduction gear and a chain transmission.

A model of this mechanism was tested at the Voronezh railroad station with satisfactory results. The flatcar is easily moved by one man who turns a handle placed at the side of the car.

The movement is easy, smooth and without jerks, at an even speed of 1.8 m/min. Since the propelling mechanism is self-braking, the flatcar comes to rest on top of the support as soon as the operator stops turning the handle of the drive. This provides a more accurate reading on the weighbridge, since the flatcar is stopped without a jerk.

The mechanism consists of a three-stage reduction gear with spur pinions and a gear ratio of 1: 32. The reduction gear operates an axle which carries a driving sprocket connected to a transmission roller-chain which drives a chain wheel, fixed to the axle of one of the flatcar wheel-pairs.

The small driving gear of the first stage of the reductor is fixed on the manually operated driving axle, whose square-shaped ends are taken outside the frame of the flatcar. The flatcar is driven by means of a single handle, fixed to the end of the axle nearest the weighing cabin.

When the flatcar is being driven as part of a train, the handle and chain are taken off and deposited in a box which is fixed between the two beam ridges of the flatcar and is closed by means of a door with a bolt. One man can dismantle and fix the chain in 7-8 min.

The mass of this mechanism (about 350 kg) is incorporated in the mass of the flatcar. The dismountable parts of the mechanism are stamped with a number which corresponds to the last two digits of the flatcar number.

In order to make a complete study and finish off the design of the above mechanism the Principal Freight Administration of the USSR Ministry of Railroads in agreement with the Committee of Standards, Measures and Measuring Instruments decided to produce and test on the South-Eastern Railroad in 1960-1961 five reference-weight flatcars supplied with the above propelling mechanisms.

MEASUREMENT BY RADIOMETRIC METHODS OF THE SPEED OF BODIES FALLING IN LIQUIDS

U. Ts. Andres, A. A. Kadushin, and G. I. Shor

Translated from Izmeritel'naya Tekhnika, No. 10, pp. 27-28, October, 1960

For many research purposes it is necessary to measure the speed of bodies falling in opaque media (colored liquids, suspensions, concentrated aerosols, and other media).

A large number of various methods and devices for measuring the speed of a ball moving through opaque liquids are described [1-5]. All these methods become unreliable when the ratio of the tube diameter to that of the ball becomes large.

In 1959 the All-Union Scientific Research Institute for the Processing of Oil and Gas developed an instrument for measuring the speed of propagation of solid bodies in liquids or of the liquid itself by means of tracer atoms, thus providing a reliable operation with large ratios of the diameter of the tube to that of the ball.

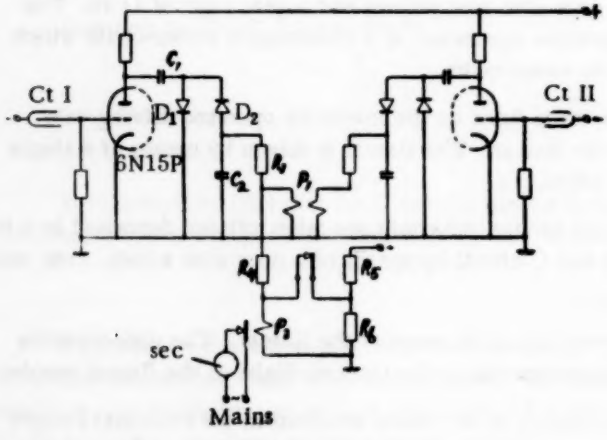

Fig. 1.

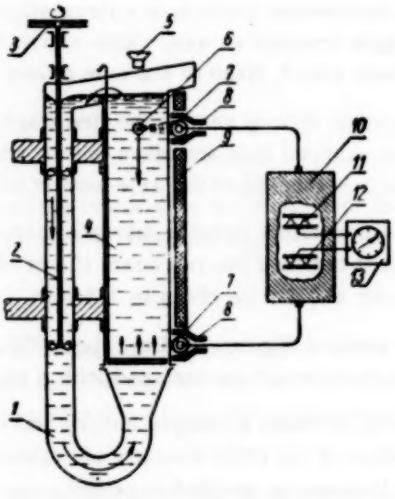

Fig. 2.

The instrument is designed for indicating (or in particular cases registering) the movement of a radioactive body in the flow under investigation. The speed of the body's movement is measured by means of two external units supplied with nuclear particle counters, which in turn register the radiations of the radioactive isotope introduced into the moving body. The external units are fixed at a distance most suitable for the given experiments. The schematic of the measuring part of the instrument is shown in Fig. 1. The radioactive radiations are received by the counter of the first external unit, producing in it negative electric pulses, which are fed to the grid of one half of tube 6N15P. The amplified pulses are fed from the anode of the tube through capacitor C_1 and a unit consisting of two diodes DGTs-12 (D_1 and D_2) to the integrating network C_2 and R_3 and the winding of relay P_1. Relay P_1 connects an electric timer through relay P_2, and locks the latter relay through a holding circuit (circuit R_4, R_5 and R_6). The timer is

disconnected by means of the second winding of relay P_1, which is connected to the integrating network of the second channel and operates when the counter of the second external unit is irradiated. Diodes D_3 and D_4 are connected in such a manner that the RC network of the second channel provides a voltage which balances out the magnetizing current of relay P_1.

In connection with the research carried out by the Institute of Fuel Minerals of the Academy of Sciences of the USSR, on the movement of bodies in viscous-plastic media (thinly dispersed mineral suspensions), the above instrument was used for measuring the speed of a sphere falling in a thinly dispersed barium-oxide aqueous suspension. Figure 2 shows the lay-out of the device for measuring the speed of the falling sphere in a rising flow of the suspension.

The equipment consists of a U-shaped tube 1, whose measuring leg is four times larger than the one containing the propeller pump 2, rotated by drive 3. Plastic balls of various densities and diameters, and supplied with radioactive cobalt cores 6, were inserted through tunnel 5 into the slowly circulating suspension, which was protected on one side by a lead screen 9. For repeated measurements the ball was recovered by means of net 4 and then reinserted. When the ball passed in front of the Geiger-Müller counter 7 in the collimated lead container 8, a pulse was produced which operated relay 10, closed the locking contact 11, and connected the electric timer 13 calibrated in 0.01 seconds. In a similar manner the relay disconnected the timer through contact 12.

The exact distance between the radiations maxima which operate the relay was determined separately for each source. In the above experiments it varied by 0.1-0.25% , depending on the activity of the source.

The maximum speed in the above experiments amounted to 20 cm/sec, and the minimum speed was 0.15 cm/sec. Each measurement was made three times, and the time intervals read off the timer did not differ from each other by more than 2%. The measured speed of a light 0.15 cm diameter ball in water (R_e = 0.85) did not differ by more than 0.5-0.7% from the value calculated by Stokes' formula.

This equipment can be used for measuring the speed of bodies moving in opaque media or as a rheoviscosimeter.

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EFFECT OF DEFORMATIONS IN THE BENDING OF THE SHAFT ON THE ACCURACY OF TORQUE MEASUREMENTS

G. A. Gulyaev

Translated from Izmeritel'naya Tekhnika, No. 10, pp. 29-30, October, 1960

It is normal practice in torque measurements to connect wire transducers to the shaft in the direction of the main stresses, i.e., at an angle of 45° to the shaft axis.

Figure 1 shows the two most widespread methods of placing the transducers and connecting them to the bridge circuit when measuring torque. With such a fixing of the transducers it was assumed that the deformations in the bending of the shaft affected the torque measurements to a negligible extent, since with this layout the effect of

bending deformations on the transducers was automatically compensated. However, the degree of compensation depends on several factors, the most important of which is the position of the transducers on the shaft being tested and the strain sensitivity of the transducers.

Often the shafts whose torque is being measured are subject to considerable bending moments, which produce deformations comparable to those due to the twisting of the shaft and sometimes even exceeding them.

In such cases the error due to the bending of the shaft may, under certain conditions, attain excessively large values.

Below we provide the results of investigating the nature of this error and the conditions of its compensation.

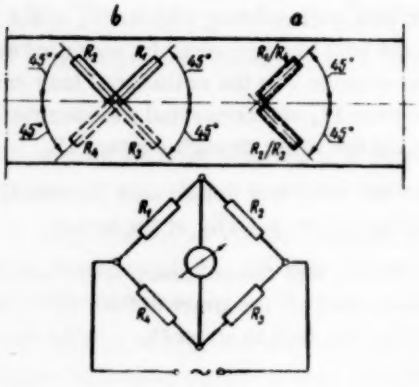

Fig. 1.

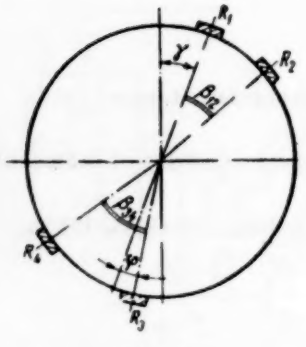

Fig. 2.

The value of the error due to the bending of the shaft depends mainly on the relative position of the transducers and the value of the torque operating on the shaft, providing the transducers have the same strain-sensitivity, attained by a careful selection of transducers with the same resistance.

The mutual position of the transducers with respect to the versions shown in Fig. 1 is determined by the following quantities (Fig. 2):

β_{12} and β_{34} are the angles between the radii passing through the centers of adjacent transducers; φ is the deviation angle of the radius passing through the third transducer's center; α_i are the angles of inclination of the transducers to the shaft axis.

Moreover, the position of the transducers on the shaft axis is determined by distance l from the center of the transducers to the datum point which we assume to be the point of application of the bending force.

An analysis of a tension position of the transducers, taken for the most common case of a cantilever action of the bending force, provided the following formula for evaluating the measurement error: *

$$\delta = \frac{2a}{\sum_1^4 l_i} [l_1 \cos\alpha_1 \cos\gamma - l_2 \cos\alpha_3 \cos(\gamma \pm \varphi) + l_4 \cos\alpha_4 \cos(\gamma + \beta_{34} \pm \varphi) - l_3 \cos\alpha_2 \cos(\gamma + \beta_{12})], \quad (1)$$

where a is the ratio of the stress in the transducer due to bending to that due to twisting; γ is the angle between the radius passing through the center of the first transducer and the direction of the bending force.

This formula has been derived with the assumption that torque is measured by means of an equal-arm bridge; moreover, it is assumed that the sensitivity of the transducers to deformations is the same.

It will be seen from (1) that the measurement error is determined by the mutual position of the transducers and the level of the bending stress, as well as by the angle of the shaft rotation with respect to the direction of the bending force. The value of the error over one turn of the shaft varies according to the cosine law.

In order to find the maximum value of the error, the extreme values of function $\delta = f(\gamma)$ must be investigated. However, investigation of this function in a general form is rather complicated, and we have, therefore,

*Formula (1) also holds for the case when the shaft can be represented as a freely supported beam, and the bending force applied at an arbitrary point. In this instance l is the distance from the center of the transducer to the end of the beam.

analyzed version b, for which $\beta_{12} = \beta_{34} = 0$, and $\varphi = 0$. The maximum of the function is then determined for $\gamma = 0$. Moreover, the measurement error thus found will be in many cases* sufficiently near to the maximum for version a of the transducer position as well.

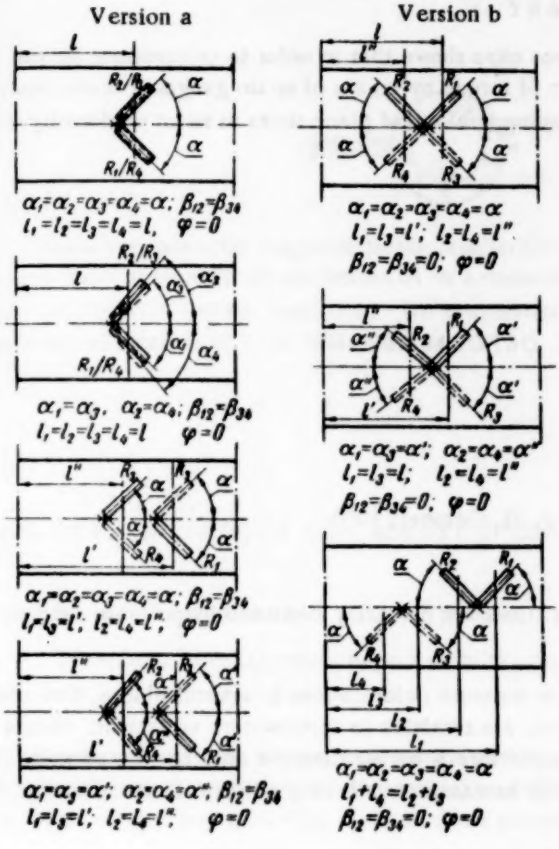

Fig. 3.

Thus the maximum measurement error is determined as

$$\delta_{max} = \frac{2a}{\sum_1^4 l_i} |l_1 \cos\alpha_1 - l_2 \cos\alpha_2 \cos\varphi + l_4 \cos\alpha_4 \cos(\beta_{34} \pm \varphi) - l_3 \cos\alpha_3 \cos\beta_{12}|. \quad (2)$$

By means of (2) conditions for a complete compensation of the measurement error can be found. These conditions with respect to the two basic methods of placing the transducers on the shaft are represented by the following equations:

Method a	Method b
1. a) $\alpha_1 = \alpha_2 = \alpha_3 = \alpha_4$;	1. a) $\alpha_1 = \alpha_2 = \alpha_3 = \alpha_4$;
b) $\alpha_1 = \alpha_3$; $\alpha_2 = \alpha_4$.	b) $\alpha_1 = \alpha_3$; $\alpha_2 = \alpha_4$.
2. a) $l_1 = l_2 = l_3 = l_4$;	2. a) $l_1 = l_3$; $l_2 = l_4$; (3)
b) $l_1 = l_3$; $l_2 = l_4$.	b) $l_1 + l_4 = l_2 + l_3$.
3. $\beta_{12} = \beta_{34}$.	3. $\beta_{12} = \beta_{34} = 0$.
4. $\varphi = 0$.	4. $\varphi = 0$.

Relations (3) provide the possibility of compiling for each basic method of fixing the transducers several versions (Fig. 3) for which complete compensation of the measurement error due to bending deformation is achieved.

It will be seen from a comparison of the two methods of fixing transducers on the shaft, that the two methods are about equally effective in eliminating the measurement error due to the bending of the shaft. However, the breaking of one of the four conditions in (3) leads to different errors in the two methods: if the condition of the equality of angles α and lengths l is broken a larger error is produced, with other conditions remaining equal, when the transducers are fixed according to method b; on the other hand, method a is more sensitive to the condition of the equality of angles β and $\varphi = 0$ being broken.

The theoretical analysis was checked on a special experimental equipment; the basic element of this equipment consisted of a steel shaft on which the strain gauges were placed according to the method under consideration.

All the possible optimum versions of fixing transducers on the shaft, as shown in Fig. 3, were tested on this equipment. All these versions provided an almost complete compensation of the bending effect; the maximum measurement error for $M_{bnd} = M_{tws}$ did not exceed according to the experiment ± 0.5-0.9%. This error is probably due to the inequality in the sensitivity of different transducers.

The nature and value of the errors when one or several equalities in (3) were broken were also investigated on this equipment.

The experimental data has also revealed a regular, periodic nature of the error, which goes through a complete period of oscillations in one turn of the shaft. Thus the error of measuring torque for each complete

*When the angles β are small or when $\alpha_2 \simeq \alpha_4$; $l_2 \simeq l_4$; $\beta_{12} \simeq \beta_{34}$; $\varphi \simeq 0$.

turn of the shaft is almost completely compensated, providing the bending moment remains constant in value and direction. The value of the maximum error obtained experimentally is in good agreement with that calculated from (2).

SUMMARY

The above experimental and theoretical investigations have shown that in order to compensate for the error produced by the bending of the shaft in measurements of torque by means of strain gauges it is necessary to fix the transducers in positions which have the smallest bending strains and place them in strict conformity with the relations deduced above.

PORTABLE EXTENSOMETER FOR MEASURING DEFORMATIONS AT HIGH TEMPERATURES

V. N. Samoilenko

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In extensive testing under working conditions, deformations are normally measured by portable extensometers.

The use of portable extensometers makes it possible to measure deformations in several places, thus making the testing simpler and cheaper. These instruments, however, are sensitive to temperature variations. Hence in measuring deformations of structures near which the air temperature is higher than the ambient temperature (for instance, when measuring thermal plants), the use of portable extensometers is only possible if the strains in the instrument due to temperature are taken into account.

The temperature expansion of the instrument is usually accounted for by two methods: the temperature of the portable extensometer is determined during testing or its dimensions are checked on a special standard before taking the reading.

Taking into account the expansion of the instrument complicates the testing and the working out of the results.

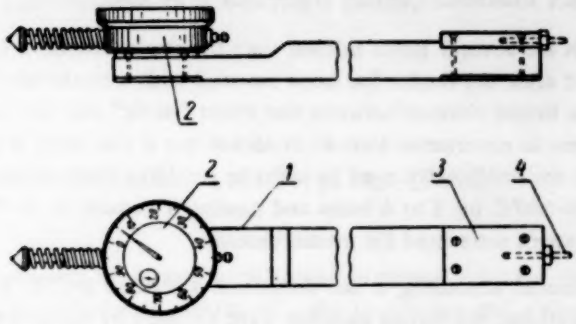
The Scientific Research Institute for Concrete and Ferro-Concrete has adopted a portable extensometer designed by the author of this article for use without the application of strain corrections.

The portable extensometer (see figure) consists of a wooden rod 1, one end of which carries a normal dial extensometer 2, and the other end a thrust plate 3, with a screw 4. The moving pin of the instrument and the thrust plate screw end in bearing balls, which are inserted into the triangular grooves of the measuring rod welded to the construction under test.

In order to avoid the effect of air humidity on the dimensions of the wooden rod it is covered with three layers of aluminum foil or three coats of carbinol cement.

The thermal conductivity of steel is about 150 times greater and its linear temperature coefficient some 2-3 times greater than that of wood. The dimensions of the wooden rod hardly vary at all during testing. Such an instrument is also lighter than an instrument with a metallic rod.

The above extensometer has been successfully used for several years in extensive testing at a surrounding air temperature amounting to 100-130° and the temperature of the construction under test to 200-270°.



Moreover, powerful magnetic fields exist in the area of measurement and the construction is under tension. The use under such difficult conditions of an extensometer with a wooden rod provided an accurate determination of the construction's strain condition. An instrument with a base of 420 mm had under the above conditions a maximum elongation of 0.02-0.04 mm.

FILM STRAIN GAUGES FOR MEASURING MAGNETOSTRICTION IN FERRITES

V. A. Yugov and G. P. D'yakov

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For the calculation of magnetostriction radiators, delay lines and similar devices, it is necessary to know the value of magnetostriction. In the past wire strain gauges [1, 2] were used for measuring magnetostriction. The layers of glue and insulation limit in these gauges [3] a further rise in the accuracy of measurement.

The successful application of film nonglued strain gauges for measuring magnetostriction in nickel [4, 5] made it possible to extend this method to the measurement of ferrites. The characteristic properties of ferrites (high thermal stability, small bulk conductivity and considerable hardness) made it possible to improve and simplify the design of the strain gauges. In designing a film strain gauge the following basic problems had to be solved: choice of material for the strain-sensitive layer; the provision of leads from the strain-sensitive layer to the measuring instrument; insuring a reliable contact between the leads and the strain-sensitive layer.

Nichrome and constantan were first used for the strain-sensitive layers. Owing to their high resistivity these layers become relatively thick (considerably exceeding the critical value) for the required value of resistance. This provided a practically continuous layer, despite the fact that the ferrite surface had irregularities of the order of 1 μ, which could not be completely removed even by careful polishing. Thus, it could be considered that the properties of the strain-sensitive layer were the same as those of a solid alloy or of wire. The strain-sensitive layers were made by evaporation in vacuum; a linear vaporizer was used for the purpose. For an improved, even distribution of the layer it is advisable to use thin tungsten and nichrome or constantan wire of uniform diameters (0.1-0.3 mm), wound on a thick tungsten wire (0.8-1.2 mm). In order to preserve the quantitative relations between the alloy components of the layer it is necessary to evaporate them at temperatures of a given range [6].

The leads to the measuring instruments were connected to fired-on silver contacts (using the normal silver carbonate paste). Conducting "heels" were deposited in the places where the ends of the strain-sensitive layers should come. Wires or narrow foils were then soldered to these silver heels, the whole surface was carefully cleaned and then covered by a strain-sensitive layer. The resistance of the layer was measured during condensation. The depositing of the film was stopped as soon as the required resistance was reached.

Experience has shown that, unless special precautions are taken, the electrical contact between the fired-on silver and the nichrome layer is insufficiently stable, its resistance sometimes exceeding that of the layer itself.

In order to avoid this effect additional contact layers had to be used.

The contact layers, which are several times thicker than the strain-sensitive layer, are also deposited by the vacuum evaporation method after the ferrite has been covered with a strain-sensitive layer. Aluminum and nickel were used for providing a better contact between the silver "heels" and the nichrome or constantan layer. It is also possible to use nichrome or constantan instead of nickel but if the layer is relatively thick it begins to peel. Completed strain-gauges are artificially aged in order to stabilize their properties. Stabilization is attained by heating them at 200-300°C for 8 to 4 hours and passing a current of 10-20 ma through them for 8 hours. After stabilization the gauges were used for measurements.

Magnetostriction was measured according to the technique described in [7]. In order to prevent the possible disturbing effect of bending vibrations, the ferrite samples were covered by strain-gauges on opposite sides. The resistance of both strain-gauges was made equal with a sufficient degree of accuracy. The strain-gauges were calibrated against reference-gauges on nickel rods and discs; moreover, the measurements obtained by means of these gauges were compared with those obtained with wire strain-gauges.

For two ferrite samples the following data were obtained:

Sample No. 7 – ferrite $MgFe_2O_4$; $\rho > 10^9$ ohm·cm. It was covered by nichrome film strain-gauges. The resistance of the gauges was 828 ohm. The value of magnetostriction was $-6.25 \cdot 10^{-6}$; I_s measurements provided a value of 85 gauss;

Sample No. 73 – ferrite $NiFe_2O_4$; $\rho_k > 2 \cdot 10^8$ ohm·cm. The sample was covered by nichrome film strain-gauges. The resistance of the gauges was 285 ohm. The striction amounted to $-11.0 \cdot 10^{-6}$; $I_s = 240$ gauss.

Even with relatively large temperature variations the film strain-gauges provided stable readings.

Owing to the fact that the sensitive layer is deposited directly onto the ferrite surface (without glue or insulation), the measurement accuracy is not affected by the errors inherent in glued-on strain-gauges when they are heated.

Since films consisting of nichrome and the recently used platinum-iridium and certain other alloys remain stable up to a temperature of 300-400°C, it became possible to study the magnetostriction of ferrite with respect to temperature.

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HIGHLY SENSITIVE PRESSURE TRANSDUCERS MADE OF ORGANIC GLASS

S. B. Stopskii

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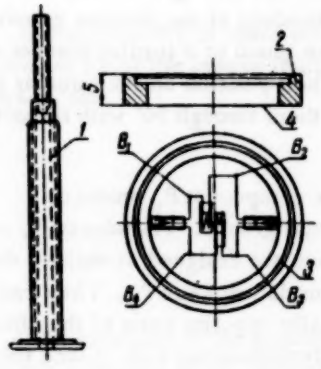
Metallic diaphragm transducers have been widely used in recent years for measuring the pressure of liquids and pulsations in their flow. The transducer consists of a diaphragm cup and a sleeve (see fig.) Four wire strain-gauges are glued to the diaphragm. Under the effect of the measured pressure the diaphragm is bent in, thus changing the resistance of the transducers. We made similar transducers from organic glass whose modulus of elasticity equals 210 kg-wt/cm^2, i.e., approximately one-hundredth of that of metal. This provides the required bending with a load amounting to tenths of the one required for a metal diaphragm of the same thickness.

When an organic glass diaphragm is used a reliable connection between the sleeve and the diaphragm is obtained and the leakage of water prevented.

The technique of making transducers from organic glass is relatively simple. The sleeve and the diaphragm cup are made of organic glass. The two operating transducers with a base of 5 mm are glued to the diaphragm, and the two compensating ones to the base of the sleeve.

Acetone-celluloid adhesive No. 11 is used for sticking the transducer. The strain-gauges are dried for 3-5 hours after gluing.



1) Transducer shell (sleeve); 2) transducer cup; 3) schematic of the positioning and connections of wire strain-gauges on the transducer diaphragm; 4) diaphragm; B_1, B_2, B_3 and B_4 are the leads of the measuring bridge.

The cups are glued to the sleeve by a 2-3% solution of organic glass in dichlorethane.

After a short drying period the joint is slightly warmed up with a soldering iron in order to insure more reliable adhesion.

No water leakage was observed in such joints.

For measuring low pressures and flow pulsations on small surfaces a small organic glass cap is fused to the transducer. A small hole is drilled at the top of this cap to coincide with the surface under investigation. The liquid is fed to the diaphragm of the transducer through this hole. The neck with the diaphragm is either screwed or glued to the surface under investigation. Such transducers are highly sensitive. We used in our investigation of pressure and pulsations in hydraulic turbine models transducers with a wall thickness of 0.4 to 1 mm.

These transducers were sensitive enough to provide pressure or rarefaction readings equal to 10 mm H_2O on a loop oscilloscope type ST-24 without an amplifier.

The transducers have a linear characteristic.

THREE-COMPONENT DYNAMOMETER

A. I. Beron, E. K. Gubenkov, and L. B. Glatman

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The cutting stresses of various materials are mostly measured by electrical dynamometers. The existing designs of dynamometers have a common defect: they have a relatively low natural frequency of oscillations, which greatly impedes investigations under conditions of self-vibrating cutters or at high frequencies of vibrations in the cutting stresses.

In the Mining Institute of the Academy of Sciences of the USSR we have developed and tested out a simple three-component dynamometer with wire transducers and devoid of these defects. The dynamometer (Fig. 1) consists of a cylindrical body 1 with stem 3 firmly connected to it, and intended for fixing to a lathe carriage or an actuating part of a machine. The sensing elements of the dynamometer consist of wire transducers, which are glued to a circular groove whose axis of symmetry passes through the middle of the cutting edge of the tool.

To register the bending deformations caused by cutting force P_z two transducers 2 of equal ohmic resistances R_1 and R_2 are glued to the circular groove in body 1 at the points of intersection of the main cross-sectional plane, which passes through the direction of force P_z and the axis of symmetry of body 1. The transducer axes coincide with the direction of the axis of the dynamometer container.

The schematic of the transducer connections is given in Fig. 2. Owing to this type of connection the bridge imbalance becomes twice as large for a given dynamometer loading by force P_z, other components of the cutting force are eliminated (P_y and P_x), and a complete temperature compensation is ensured.

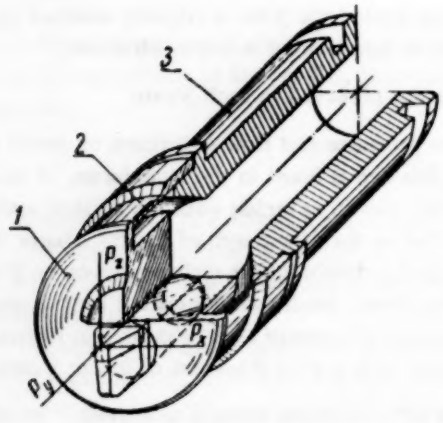

Fig. 1.

In order to register the cutting force component P_x, which produces a side bending of the circular groove, transducers R_5 and R_6 are glued in a similar manner as for measuring P_z, but their position on the circular groove is changed by rotating them through 90° with respect to transducers R_1 and R_2.

The cutting force component P_y produces a contraction of the circular groove. Transducers R_3 and R_4 of equal ohmic resistance are designed to register the contraction deformations due to force P_y. They can be glued in any diametrically opposite parts of the circular groove parallel to the dynamometer axis. Since these transducers are connected in series to one arm of the measuring bridge, all distortions due to deformations of the circular groove by cutting forces P_z and P_x are completely eliminated.

In order to avoid the possible effect of temperature variations on the ohmic resistance of the measuring transducers R_3 and R_4, compensating transducers K_1 and K_2 of equal ohmic resistance with the measuring transducers are placed at right angles to each of them. The compensating transducers are also connected in series and form the second arm of the measuring bridge.

The balancing of the measuring bridge arms of P_z, P_y and P_x is attained by bending the corresponding balancing rods of the strain-amplifier to which similar transducers to the measuring ones are glued. They constitute the remaining two arms of the measuring bridge circuit.

The sensitivity of the measuring arm P_y is smaller than that of arms P_z and P_x. It can be raised, however, if the balancing of the measuring system of transducer deformations of the remote half-bridge by means of bending the balancing rod is dispensed with, and all the measuring and compensating transducers are glued to the measuring groove of the dynamometer body.

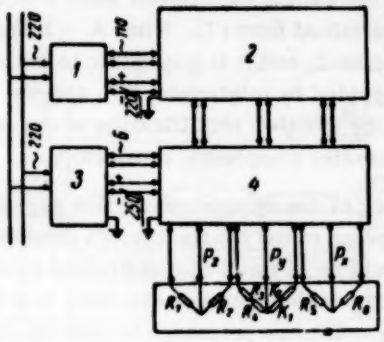

Fig. 2. 1, 3) Rectifiers; 2) oscilloscope; 4) amplifier.

The selection of the design dimensions of the dynamometer body and its thickness in the area of the groove to which the transducers are glued is affected by the probable relation of the cutting forces P_z and P_y, as well as the amount by which the cutter protrudes beyond the line on which the transducers are glued in the groove. The method of fixing the cutter in the body of the dynamometer (due to constructional considerations) determines the amount by which the cutter protrudes beyond the circular groove H, and the diameter and thickness of the groove are selected on the basis of the following considerations.

The stress in the body of the dynamometer due to the application of force P_z at the place where the transducers are glued is equal to

$$\sigma_z = \frac{32HDP_z}{\pi (D^4 - d^4)}, \tag{1}$$

and the compression stress of component P_y is

$$\sigma_y = \frac{4P_y}{\pi (D^2 - d^2)}, \tag{2}$$

where D is the external diameter of the circular groove, mm; d is the internal diameter of the circular groove, mm.

In the majority of cases thickness δ of the dynamometer wall at the place where the groove is made doe not exceed 5% of the groove diameter, and therefore it is possible to consider, without greatly affecting the accuracy, that

$$\sigma_z = \frac{4HP_z}{\pi D^2 \delta}, \tag{3}$$

$$\sigma_y = \frac{P_y}{\pi D \delta}. \tag{4}$$

The selection of the groove diameter D and the thickness of the wall δ is aimed at obtaining the greatest permissible tension in the dynamometer body at the places where the transducers are glued. Considering that during the cutting process tensions arise in the groove due simultaneously to forces P_z and $P_y = \lambda P_z$ we have

$$\delta = \frac{P_z}{\pi D \sigma_{com}} \left(\lambda + \frac{4H}{D} \right). \tag{5}$$

For the purpose of the best possible utilization of the dynamometer material and obtaining uniform sensitivity of the measuring elements of all the components, it is desirable to make tensions δ_z and δ_y equal. From the condition of the equality of these tensions it follows that

$$\frac{D^2 + d^2}{DH} = \frac{8}{\lambda}. \tag{6}$$

When the thickness of the dynamometer body at the place of the groove does not exceed 5% of the groove diameter it is possible to assume, without affecting the accuracy of calculations, that in (6) $D \approx d$. Then

$$\frac{D}{H} = \frac{4}{\lambda}. \tag{7}$$

In constructing three-component dynamometers for investigating the cutting of minerals when $\lambda > 1$, satisfactory results are obtained if the dynamometer dimensions are determined from (7). When $\lambda < 1$ (for instance, in cutting metals), the dimensions obtained from (7) cannot be used, and it is impossible to maintain the condition of the equality of σ_z and σ_y. In this case one should be guided by relationship (5), and the slightly reduced sensitivity of measuring bridge P_y can be compensated by a higher amplification or the use of a more sensitive loop oscilloscope. The accuracy of measuring P_y still remains completely satisfactory.

In order to provide the highest possible natural resonance frequency of the dynamometer, it is necessary to make the thickness of the groove as large as possible (but not at the expense of the dynamometer's sensitivity), but the mass of the part of the dynamometer which holds the cutter should be as small as possible and should be placed as near as possible to the center of the circular groove. The fixing of the dynamometer body to a lathe carriage or an actuating element must be sufficiently rigid and with the minimum protrusion beyond the supporting surfaces.

The technical characteristics of the manufactured and successfully tested dynamometer are as follows: calculated cutting force P_z = 3000 kg-wt, P_y = 12,000 kg-wt, P_x = 3,000 kg-wt; natural frequency of oscillation in the directions P_z and P_x is 1,200 cps; the diameter and thickness of the circular groove are respectively 10.0 and 0.25 cm; the cutter protrudes beyond the axis of the circular groove by 7 cm; the resistance of the wire transducers is 200 ohm; eight transducers were used.

The calibration of the dynamometer with a static load provided a strictly linear scale. The reading error of the dynamometer equipment amounts to 1.0% of the nominal load, and in the range of 0.2 to 1.0 of the nominal load the error does not exceed 2.0% of the measured value.

The stability of the dynamometer readings has been checked by 800 to 1,200 experiments in cutting minerals. In all these instances the calibration charts remained the same.

PRODUCTION OF WIRE TRANSDUCERS

N. N. Butorin

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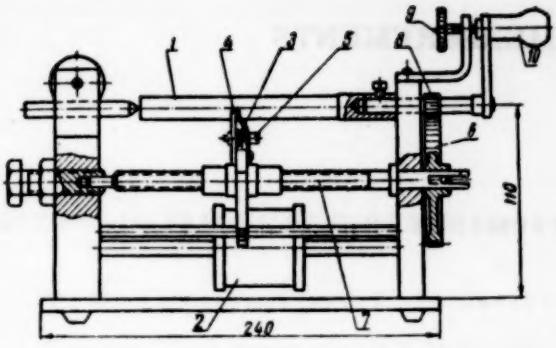
A simple lathe for making wire transducers is being successfully used in the Arkhangel' Forestry Institute. The following are the advantages of the above method: double-layer transducers are made which have an ohmic resistance and power twice as large as single-layer transducers of the same size; the transducers have sharp loops and therefore their sensitivity to transverse deformations is considerably lower than in normal transducers with circular loops; the productivity of the lathe is high, since ten transducers can be wound on a single paper former.

The construction of the lathe is shown in the figure attached. A strip of cigarette paper is wound once around an exchangeable spindle 1, and fastened to it by means of three rubber rings. The width of the strip is selected to make its edges overlap each other by 1-2 mm. The diameter of the spindle is selected to fit the required dimension of the transducer base.

The constantan wire is drawn from coil 2 over pulley 3 and under the lathe carrier spring 4, which is adjusted by means of screw 5 for tensioning the wire.

The gear wheel 6, fixed to the bracket of guide screw 7, is displaced along the shaft and disengaged from the small gear 8. Gear wheel 6 is then rotated together with the guide screw until lathe carriage 4 is placed against the right-hand corner of the paper. The end of the wire is then placed on the paper and glued to it. As soon as the glue dries, gear wheel 6 is engaged with gear 8, and stop 9 is unscrewed, freeing handle 10. As the handle is turned the wire is wound in even rows over the paper backing. The spacing between the turns for a gear ratio of 6 and a guide-screw pitch of 1.5 mm is equal to 0.25 mm. Having wound the required number of



turns for a given ohmic resistance of the transducer (about 15-20 turns), the last turn is fixed by means of gluing. The handle is then secured by the stopping device, the gearing disengaged and, having displaced the lathe carriage to the left by a few millimeters, the winding of the next transducer is started.

When all the transducers have been wound they are covered with glue and taken off the lathe. As soon as the glue dries the paper tube with the transducers is taken off the spindle and cut by means of scissors into separate rings with one transducer to each ring. These rings are then smeared inside with glue and placed under a press. After drying the transducers, lead-out wires are soldered to them and cigarette paper is glued to them on both sides.

THERMOTECHNICAL MEASUREMENTS

APPLICATION OF FREON FOR CHECKING INSTRUMENTS IN THE LOW TEMPERATURE RANGE

A. D. Brodskii and G. N. Filandrov

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For state testing of various thermometers and thermocouples in the 0 to −100°C temperature range, alcohol cryostats with liquid oxygen or nitrogen cooling are normally used. A uniform temperature field inside the bath is usually obtained by stirring the alcohol with a mixer, and a temperature equilibrium is obtained by a simultaneous control of cooling and electric heating. In liquid bath cryostats it is normal to use for temperatures below −100°C butane-propane (solidifying temperature − 180°C) or pentane (−200°C) instead of alcohol. It is, however, difficult to work with these gases since they are explosive. In this connection the Low Temperature Laboratory of the VNIIM (All-Union Scientific Research Institute of Metrology) has examined the possibility of using for testing purposes the chemically inert and explosion-proof gas freon. The table below shows the technical data of certain freon gases produced by our chemical industry.

Freons	Conventional designation	Temperature, °C	
		of boiling	of melting
Freon-12	F-12	−29.8	−155.0
Freon-13	F-13	−81.5	−180.0
Freon-14	F-14	−127.9	−184.0
Freon-22	F-22	−40.8	−160.0

Using Freon-12 as a cooling agent, we compared P. G. Strelkov's reference resistance thermometers in an ordinary Hemming type cryostat at temperatures of −41.9; −62.6; −69.6; −105.9; and −131.8°C. The constancy of the maintained temperature in the cryostat amounted to 0.01-0.02 deg/min. The horizontal gradients were 0.007-0.040 deg/cm, and the vertical ones 0.008-0.050 deg/cm. Prolonged storage of freon was accomplished by cooling it periodically down to the temperature of −155°C. In our experiment freon was preserved for 15 days.

In case special cryostats are not available, thermometers and thermocouples can be checked in ordinary Dewar flasks with a firmly fitting cap. The checking is then made at the boiling and melting points of various types of freon. By mixing various types of freon it is possible to obtain liquids which boil at a given temperature. It should be noted that an open flame should never by approached to freon, since phosgene is then formed.

The melting and boiling temperatures of freon are determined by means of a reference resistance-thermometer or a standard thermocouple. It should be noted that at the present time freon is a relatively expensive material; however, the replacement of alcohol by freon in mass-production testing makes it possible to extend the calibration range of instruments and simplify the testing process itself.

SYSTEMATIC ERRORS IN DIGITAL TIME-PULSE CONVERTERS

V. N. Khlistunov and V. P. Lavrov

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Various methods of converting analogue quantities into a digital form are used in digital voltmeters and ammeters. However, questions related to the errors of conversion have not received the required attention in literature. The present article deals with the investigation of systematic errors in time-pulse converters.

The basic error components can be divided into three groups: errors due to the instability of supplies; errors due to variations of ambient temperature; and errors due to the use of discrete pulses.

The latter error was not examined separately in this work and its value was taken to be ±1 pulse, which is always the case if special measures for synchronizing the operation of discrete elements of the circuit are not taken. Errors were analyzed for converter circuits using voltage generators with linearly-falling characteristics and one diode comparing device. The converted (measured) voltage U_X is fed to the generator anode (Fig. 1) through a clamping diode. The above conversion circuit is most suitable for measuring rapidly changing quantities, since it does not require the use of an input device which clamps the instantaneous values of the measured variable, and since it provides a very simple registration of their values as functions of time. A model of such a converter was checked experimentally.

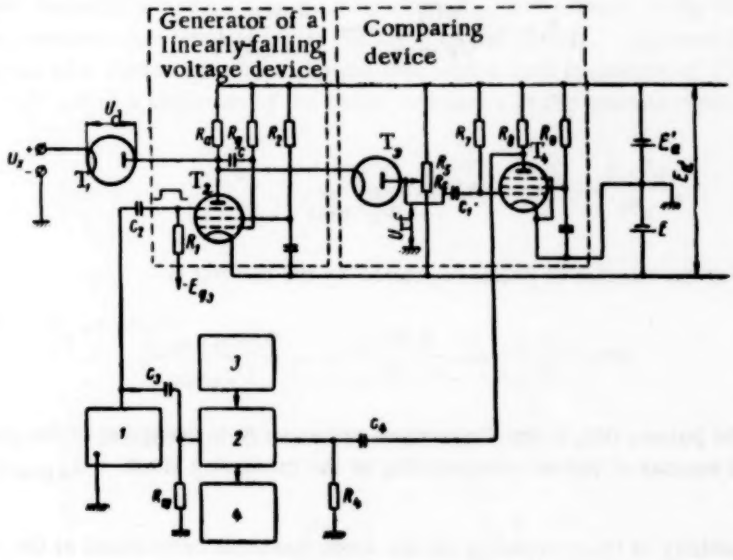

Fig. 1.

The principle of operation of this circuit consists of the following. At the instant measurement begins the starting circuit 1 sends a long control pulse to the suppressor grid of the linearly-falling-voltage generator tube T_2 and a short pulse to the electronic switch 2. Under the effect of the controlling pulse the anode voltage of the generator begins to fall according to a linear law from its initial value U_X to the value of the reference voltage U_{rf}, which is close to ground potential, and the conducting electronic switch transmits the generator count pulses 3 of a reference frequency to the electronic pulse-counter 4.

When the generator anode voltage reaches the value of U_{tf}, diode T_3 of the comparing device becomes conducting and the control grid of the amplifier-limiter tube T_4 receives a signal which blocks the electronic switch through a feedback circuit.

The number of count pulses registered by the counter in the interval between two controlling pulses, produced by the electronic switch, is directly proportional to the voltage U_X.

ERRORS DUE TO INSTABILITY OF SUPPLY VOLTAGE

Effect of variations in the heater voltage of the tubes. Errors due to the instability of the heater voltage arise in the following elements of the circuit: the clamping diode T_1; the comparing device; the generator of the linearly-falling voltage.

Error due to the instability of the falling voltage in the clamping diode. A variation in the heater voltage of the diode produces a displacement of its volt-ampere characteristic, thus leading to a variation in the voltage drop across it. This leads to a variation in the initial voltage across the generator anode for a given U_X.

The voltage with respect to ground at the generator anode is.

$$U_a = U_X + U_d,$$

where U_d is the voltage drop across the clamping diode.

According to the published data, a variation in the diode heater voltage of ±10% produces a parallel displacement of the diode characteristic amounting to $\mp$ 0.1 v.

If we assume a linear relationship between the displacement of the volt-ampere characteristic and the variation of the heater voltage we may assume that

$$\Delta U_2 = -0.01\gamma_1, \tag{1}$$

where ΔU_2 is the displacement of the volt-ampere characteristic, v; γ_1 is the variation in the heater voltage, %.

The above mentioned error due to the instability of the diode heater voltage will now be expressed as

$$\delta_1 = -\frac{\Delta U_2}{U_{x\,\max}} = -\frac{0.01\gamma_1}{U_{x\,\max}}, \tag{2}$$

or in percentage

$$\delta_1 = -\frac{\gamma_1}{U_{x\,\max}}. \tag{3}$$

The error expressed as the number of pulses registered by the counter is

$$dn_1 = \frac{\Delta U_2}{\Delta U_1} = -\frac{0.01\gamma_1 f_0}{K} = -\frac{0.01\gamma_1 n_{\text{nom}}}{U_{x\,\max}}, \tag{4}$$

where dn_1 is the error in the pulses; ΔU_1 is the discreteness value, v; K is the speed of the generator voltage variation; n_{nom} is the nominal number of pulses corresponding to the measuring limit $U_{X\,max}$; f_0 is the repetition frequency of count pulses.

Error due to the instability of the comparing device diode heater is determined as the variation in the blocking voltage of the diode.

The above error in this instance will be

$$\delta_2 = \frac{\Delta U_2}{U_{x\,\max}},$$

where ΔU_2 is the variation in the diode blocking voltage.

This variation is determined from (1). In percentage it will be

$$\delta_2 = \frac{\Delta U_2 \cdot 100}{U_{x\,\max}} = -\frac{\gamma_1}{U_{x\,\max}}.$$

Thus the errors δ_1 and δ_2 have similar signs and are added arithmetically.

Error due to the instability of the heating voltage of the linear-falling-voltage generator. Variations in the heater voltage of the generator tube of $\pm\gamma_1 = \pm 10\%$ lead to a relative error of $\delta_3 = \mp 0.04\%$ [1].

Assuming a linear relationship between δ_3 and δ_1 in the range of $\gamma_1 = 0\text{-}10\%$, we obtain in a general form

$$\delta_3 = -0.004\,\gamma_1. \tag{5}$$

The error expressed in pulses is equal to

$$dn_3 = \frac{\delta_3 n_{\text{nom}}}{100} = -\frac{0.004\,\gamma_1}{100} n_{\text{nom}}. \tag{6}$$

Effect of anode voltage variations. Errors due to the instability of the anode voltage arise in the same circuit elements as those due to the heater-voltage variations.

Error in the clamping diode (Fig. 1). The voltage at the anode of the linearly-falling-voltage generator during the period preceding its operation is represented by the formula

$$U_a = U_x + U_d = U_x + \frac{E'_a - U_x}{R_a + R_d} R_d, \tag{7}$$

where R_d is the dc resistance of the clamping diode.

Since the anode load consists of a very large resistance $R_a[R_a \gg R_d]$, the nonlinearity of the diode characteristic can be neglected, assuming that $R_d = R_{dn} = \text{const}$, where R_{dn} is the value of the diode resistance when the nominal value of the measured resistance is fed to the input of the circuit. Resistance R_{dn} is of the order of 1 kohm. Tests carried out with diodes type 6Kh6 showed that with a load resistance of 1-2 meg variations of the anode voltage of 25-30% led to a variation of the anode resistance at the operating point of 2-3%.

$$U_a = U_x + \frac{E'_a R_d}{R_a + R_d} - \frac{U_x R_d}{R_a + R_d} \simeq U_x + \frac{E'_a R_d}{R_a} - \frac{U_x R_d}{R_a}, \tag{8}$$

since $R_d \ll R_a$.

Denoting $R_d/R_a = \Theta$, we obtain

$$U_a = U_x + E'_a \Theta - U_x \Theta. \tag{9}$$

Denoting $\frac{dE'_a}{E'_a} 100\% = \gamma_2$, we obtain the referred error in percentage:

$$\delta_4 = \frac{dU_a}{U_{x\,\max}} 100 = \frac{E'_a \Theta}{U_{x\,\max}} \gamma_2. \tag{10}$$

The error expressed in the number of pulses is

$$dn_4 = \frac{E'_a \Theta \gamma_2}{U_{x\,\max} \cdot 100} n_{\text{nom}}. \tag{11}$$

Error due to the effect of the anode voltage variations on the reference voltage. Here we examine a case when the reference voltage U_{rf} is obtained from a common anode voltage by means of a potential divider R_5, R_6. The equivalent circuit of the reference voltage is shown in Fig. 2.

Let us denote

$$\frac{R_5}{R_5+R_6}=\theta_1;$$
$$\frac{R_6}{R_5+R_6}=\theta_2; \tag{12}$$

then

$$U_{rf}=E'_a\theta_2-E\theta_1. \tag{13}$$

Let us now find the variation in the reference voltage due to the variation of the supply voltage:

$$dU_{rf}=\theta_2 dE'_a-\theta_1 dE. \tag{14}$$

Let us denote

$$\frac{dE'_a}{E'_a}\,100\%=\gamma_2;$$
$$\frac{dE}{E}\,100\%=\gamma'_2,$$

and find the referred error in percentage:

$$\delta_5=\frac{\theta_2\gamma_2E'_a}{U_x\ \max}-\frac{\theta_1\gamma'_2E}{U_x\ \max}. \tag{15}$$

For $\gamma_2=\gamma'_2$, which nearly always is the case, we have

$$\delta_5=\frac{\gamma_2}{U_x\ \max}(E'_a\theta_2-E\theta_1)=\frac{\gamma_2}{U_x\ \max}U_{rf}. \tag{16}$$

The error expressed in the number of pulses is

$$dn_5=\frac{\gamma_2(E'_a\theta_2-E\theta_1)n_{nom}}{100U_x\ \max}=\frac{\gamma_2U_{rf}n_{nom}}{100\,U_x\ nom}, \tag{17}$$

where n_{nom} is the number of pulses for a nominal value of E'_a and E.

It will be seen from the last two formulas that for $U_{rf}=0$ there will be no error of this type, i.e., variations of the anode voltage will not produce variations in the reference voltage. On the basis of the tests made on the digital voltmeter model for R_5 = 200 kohm, R_6 = 110 kohm (U_{rf} is close to zero) and $\gamma_2=\pm2\%$ Table 1 was compiled.

The data of Table 1 were obtained from repeated measurements at each point (20 measurements) by means of a digital instrument with a discreteness error of ± 1 pulse (0.1%). Of the series of measurements obtained for each point the predominant one was taken as the correct result. The anode voltage of the remaining circuit was maintained constant with the accuracy of potentiometer type PPTV-1.

From the analysis of the results given in Table 1 it is possible to assert, with an accuracy of the discreteness of the model, that no errors were observed due to the variation of the reference voltage.

TABLE 1

U_x, V \ U sup, V	E_a (435 V)	E_a +1%	E_a +2%	E_a −1%	E_a −2%
100	1000	1000	1001	1000	1000
50	500	499	501	500	500
25	250	251	250	500	249
0	1	0	1	1	1

TABLE 2

U_x, V \ U sup, V	E_a (435 V)	E_a +1%	E_a + 2%	E_a −1%	E_a −2%
100	1000	992	981	1013	1021
50	500	496	491	507	511
25	250	250	246	254	256
0	1	1	1	1	1

Errors of the linearly-falling-voltage generator (error due to the variations in the control grid supply voltage). If the control grid of the generator tube is fed from a separate source E_g, the conversion time will be expressed by the formula

$$t=CR_g\frac{U_x}{E_g}.$$

In the circuit under consideration the grid is fed from a common anode supply; hence, the conversion time will be

$$t=CR_g\frac{U_x}{E_a},$$

where E_a is the generator anode supply voltage.

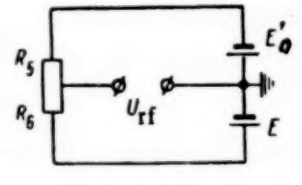

Fig. 2.

Let us assume that the supply voltage E_a is maintained constant on the electrodes of the remaining tubes.

Denoting the relative percentage variation of the anode voltage by γ_2, we shall obtain an expression for the absolute error in the conversion time:

$$dt\simeq-\frac{CR_g U_x}{100\,E_a}\gamma_2=-\frac{t\gamma_2}{100}. \tag{18}$$

The relative error in percentage is

$$\delta_c=\frac{dt}{t}100=-\gamma_2. \tag{19}$$

Thus, the relative error due to the generator control-grid voltage variations is equal to the relative variations of the anode supply voltage, taken with a reversed sign. This error expressed in the number of pulses is

$$dn_6 = \frac{\gamma_2 n_{nom}}{100}. \quad (20)$$

The experimental results given in Table 2 are in good agreement with the calculations.

The reference voltage in this experiment was maintained constant with the accuracy of potentiometer type PPTV-1.

The error due to voltage variations on the screen grid partly includes the instability due to the initial jump of the anode voltage of the linearly-falling-voltage generator, since the variation in the voltage of the screen grid leads both to a different speed in the generator voltage variation and to a change in the initial jump.

It is known that in order to maintain a linear variation of the voltage across the capacitance of the time-setting anode circuit of a linear-falling-voltage generator, the current of the tube must rise according to a linear law. This is attained by establishing between the anode and control grid of the tube an appropriate feedback which provides with a decreasing voltage across the capacitance a corresponding rise in the control grid voltage.

Owing to the negative anode current feedback, variations of the screened grid supply voltage will produce changes in the control grid voltage. The value of these variations can be obtained from expression

$$\Delta U_{gc} S_c = \Delta U_{gs} S_s, \quad (21)$$

where ΔU_{gc} is the voltage variation on the control grid; S_c and S_s are the slopes of the anode characteristics with respect to the control and screened grids respectively; ΔU_{gs} is the voltage variation on the screened grid.

Relationship (21) expresses the effect of the anode current negative feedback of the tube.

It follows from the above that the error due to the voltage variation across the screened grid can be evaluated on the basis of the relationships already obtained in determining the error due to the instability of the control grid voltage supply. Since the screened grid voltage is obtained from the common anode voltage supply, it is only necessary to take into account when using (19) the effect on the anode current of equal voltage variations on the control and screened grids (21).

The effect of control grid voltage variation is 15-30 times greater (depending on the type of the tube and operating conditions) than that of the screened grid. Hence

$$\frac{\delta_6}{\delta_7} = \frac{S_c}{S_s}.$$

where δ_6 is the relative error due to the control grid voltage supply variations; and δ_7 is the relative error due to the screened grid supply voltage variation.

The relative error due to the screened grid supply voltage variation in percentage is

$$\delta_7 = \frac{S_s}{S_c}\delta_6 = \frac{S_s}{S_c}\gamma_2. \quad (22)$$

For $S_s / S_c = 0.05$

$$\delta_7 = 0.05\ \gamma_2.$$

The error expressed in the number of pulses is

$$dn_7 = \frac{0.05\, n_{nom}}{100}\gamma_2. \quad (23)$$

The data thus obtained are in good agreement with the experimental results given in [1].

Total error due to variation of the supply voltage. The total referred error due to the instability of the sources of supply is

$$\delta_\Sigma = -\frac{\gamma_1}{U_x \max} - \frac{\gamma_1}{U_x \max} - 0.004\gamma_1 + \frac{E'_a \gamma_2}{U_x \max}\Theta + \frac{\gamma_3}{U_x \max}(E'_a\Theta_2 - E\Theta_1) - \gamma_2 - 0.05\gamma_3. \quad (24)$$

The tested model had the following characteristics: E'_a = 285 v; E = 150 v; $U_{x\max}$ = 100 v; Θ = 0.0013; Θ_1 = 0.65; Θ_2 = 0.35.

Moreover,

$$\delta_\Sigma = -0.024\gamma_1 - 1.05\gamma_3, \quad (25)$$

or in pulses

$$dn = \frac{0.024\gamma_1 n_{nom}}{100} - \frac{1.05\gamma_3 n_{nom}}{100}. \quad (26)$$

The relationships thus obtained are in good agreement with experimental data.

The model was also tested for stability of readings with time. For this experiment an anode supply source was used with a total instability (including a "drift") of ± 0.1%. The heater voltage was not stabilized but it was continuously checked by a grade 0.5 instrument. Heater voltage variations exceeding 5% were not observed.

The anode voltage supply was controlled by means of potentiometer type PPTV-1.

The error of the tested model which had n_{nom} = 1,000 amounted to ±2 pulses (Table 3) in 8 hours of continuous operation. The calculated error determined for the worst conditions observed during the testing with γ_1 = ±5% and γ_2 = ±0.1% and calculated from (25) was equal to

$$\delta_\Sigma = \mp 0.024 \cdot 5 \mp 1.05 \cdot 0.1 \simeq \mp 0.225\%.$$

The error obtained in another model with $n_{nom} = 10^4$, and a total instability of the anode supply of ±0.02% and in the heater voltage of ±1%, amounted to 4-5 pulses.

The calculated error for this case was δ_Σ = 0.024 · 1 + 1.05 · 0.02 = 0.045%, or 4-5 pulses.

ERRORS DUE TO VARIATION IN AMBIENT AIR TEMPERATURE

Errors of this type are mainly due to variations of the parameters of the time-setting circuit $C-R_g$ and the parameters of the reference frequency generator tuned circuit.

Errors due to the variations of the time-setting circuit $C-R_g$ caused by temperature variations.

The relation of the capacitance to temperature has the form

$$C = C_0[1 + \alpha_C(t° - 20°)], \quad (27)$$

where C_0 is the capacitance value at t° = 20°C; α_C is the temperature coefficient of the capacitance.

The relation of the resistance to temperature is expressed by the formula

$$R_g = R_0[1 + \alpha_R(t° - 20°)], \quad (28)$$

where R_0 is the value of the resistance at t° = 20°C; α_R is the temperature coefficient of the resistance.

Coefficients α_C and α_R can be either positive or negative.

The measurement time is

$$t = CR_g\frac{U_x}{E_a} = C_0R_0[1 + \alpha_C(t° - 20°)] \times [1 + \alpha_R(t° - 20°)]\frac{U_x}{E_a} \quad (29)$$

$$\frac{dt}{dt^\circ} = C_0 R_0 \frac{U_x}{E_a} (\alpha_C + \alpha_R + \alpha_C \alpha_R). \tag{30}$$

Denoting the relative percentage variation of temperature (with respect to a nominal temperature of t° = = 20°C) by γ, we obtain

$$\frac{dt^\circ}{20^\circ} 100\% = \gamma, \tag{31}$$

whence

$$dt^\circ = \frac{\gamma \cdot 20^\circ}{100}.$$

Then the absolute time error will be

$$dt = \frac{C_0 R_0 U_x}{100 E_a} 20^\circ \gamma \ (\alpha_C + \alpha_R + \alpha_C \alpha_R) = \frac{t \cdot 20^\circ \gamma}{100} (\alpha_C + \alpha_R + \alpha_C \alpha_R). \tag{32}$$

TABLE 3

Measured voltage, v	Required readings	Instrument readings	Maximum referred error, %
100	1000	999-1002	+0.2
80	800	798- 801	−0.2
50	500	498- 501	−0.2
25	250	248- 251	−0.2
10	100	99- 100	−0.1
5	50	49- 50	−0.1
1	10	9- 12	+0.2

It will be seen from (32) that the maximum error for t_{max}, which corresponds to the limit of measurements $U_{x\,max}$, amounts to

$$dt_{max} = \frac{t_{max} \cdot 20^\circ \gamma}{100} (\alpha_C + \alpha_R + \alpha_C \alpha_R). \tag{33}$$

The referred error is

$$\delta = \frac{dt_{max}}{t_{max}} 100\% = 20^\circ \gamma \ (\alpha_C + \alpha_R + \alpha_C \alpha_R). \tag{34}$$

Since $\alpha_R + \alpha_C + \alpha_C \alpha_R \ll 1$, the error due to a relative percentage temperature variation of γ will be less than γ by a factor of

$$\frac{1}{(\alpha_C + \alpha_R + \alpha_C \alpha_R) \ 20^\circ}.$$

If the temperature coefficients α_C and α_R are chosen in such a manner that $\alpha_C + \alpha_R + \alpha_C \alpha_R = 0$, the temperature error due to the parameter variations of circuit $C-R_g$ will be eliminated. In practice the temperature error will be eliminated for the condition of $\alpha_C + \alpha_R = 0$, since

$$\alpha_C \alpha_R \ll \alpha_R \text{ and } \alpha_C \alpha_R \ll \alpha_C.$$

It follows from the above that α_C and α_R should be made as small as possible and in an ideal case should be equal in magnitude and of opposite sign. A good combination is provided by using for the capacitance in circuit $C-R_g$ a mica capacitor of group G with $\alpha_C = \pm 50 \cdot 10^{-6}$ 1/deg and a manganin resistance with $\alpha_R = \mp 30 \cdot 10^{-6}$ 1/deg.

Error due to variations in the reference frequency with temperature. The reference frequency stability with temperature variations is determined in the main by the temperature stability of the quartz crystal resonator.

Normally the temperature coefficients of quartz crystal resonators have the value of the order of $\alpha = (10^{-6}$ to $10^{-6})$ 1/deg. Then we have

$$\frac{df_0}{f_0} = \alpha\,(t^\circ - 20^\circ). \tag{35}$$

The error due to frequency variations produced by temperature changes and expressed in the number of pulses is

$$dn = df_0 t = f_0 \alpha\,(t^\circ - 20^\circ)\,t, \tag{36}$$

where t is the time interval corresponding to a value of U_X.

The relative error produced by the instability of the generator frequency does not depend on the value of the measured voltage or its frequency.

SUMMARY

1. The largest component of the error produced by the supply voltage variations is that due to the linearly-falling-voltage generator caused by anode voltage changes.

2. The error due to anode voltage variations is tens of times greater than that due to the same percentage variation of the heater voltage.

3. The error due to ambient temperature variations over wide ranges is very small and can always be reduced to the required minimum.

4. The outlined analytical technique can be applied to any time-pulse conversion circuits.

LITERATURE CITED

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CIRCUITS FOR SORTING ELECTRICAL COMPONENTS

V. A. Dvinskikh

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Circuits for measuring the slope of the grid-anode characteristic of amplifying electronic tubes and the impedance of capacitors and induction coils are rather complicated. In order to avoid the effect of pick-up on measuring results, the mains frequency is not used for measuring purposes. Standard frequencies adopted for measuring the slope of the anode characteristic lie in the range of 400 to 1,600 cps. For measuring small capacitances and inductances frequencies of the order of hundreds of kilocycles are used. It is, therefore, necessary to include in the circuit an oscillator which would provide the above frequencies. Moreover, owing to the type of measurements used the signals are weak and have to be amplified. Hence, in the majority of circuits

appropriate amplifiers are included. Moreover, equipment designed for measuring the above parameters automatically includes not only complicated measuring circuits, but also units for automatically determining the measuring limits and tuning the required circuits. Thus the design becomes rather complicated.

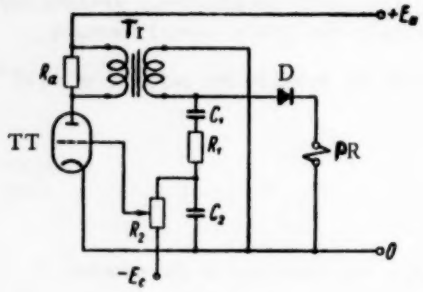

Fig. 1. TT is the tube under test; R_a is the calibrated resistance in the anode circuit; Tr is the isolating transformer; R_1, C_1, R_2 and C_2 constitute a twin RC network which transmits signals of frequency ω_0 without changing their phase; D is a detector; PR is a polarized relay.

There exists another method of designing circuits for sorting electrical components [1, 2] which is based on a combination of the functions of an amplifier and oscillator and provides a simpler circuit design.

The circuit for sorting triodes by their anode slope characteristic (Fig. 1) consists of an amplifier with a positive feedback (with an appropriate connection of the transformer windings). Obviously, if the gain of the tube under test

$$K = SR_a \qquad (1)$$

(for $R_a \ll R_i$, where R_i is the internal anode impedance of the tube) exceeds at frequency ω_0 the attenuation of the feedback circuit, continuous oscillations of frequency ω_0 will arise in the circuit. By choosing R_a in such a manner that oscillations arise only in tubes whose slope is greater than a given value, we can divide the tubes under test into two groups: the first with a slope of the anode characteristic greater than a given value, and the second with the slope smaller than that value. The oscillations thus generated are detected and operate a relay, which connects an actuating mechanism and signals "good." If the relay does not operate the tube is rejected.

In production of amplifying electronic tubes two sorting limits are taken into consideration: the lower limit examined by us, and an upper limit which serves the manufacturing plant to reject tubes with a slope of the anode characteristic higher than that limit. In this case the value of R_a is adjusted to produce oscillations when the higher limit is exceeded and the operation of the relay is made to reject tubes.

It should be noted that, if the working conditions of the tested tubes are set to specification, the tubes will operate under soft oscillatory conditions and their sorting will correspond to their actual characteristics. However, for working conditions distinct from the above, or for certain types of tubes, oscillations may arise under hard oscillatory conditions, for which the relay may operate on external shocks with tubes whose slopes correspond to small grid excitation voltages below the set limits. In order to eliminate this defect it is necessary first to set the specified working conditions for the tube, and then close the feedback circuit through the grid circuit and insure that this connection is not broken during testing. All these conditions are easily satisfied by installations of the turret or conveyor type which have a pulsating type of movement.

Exhaustive tests of the above equipment for sorting tubes type 1K2P, which have a lower limit for the slope of their anode characteristic of 0.30 ma/v for $R_1 = R_2 = 100$ kohm, $C_1 = C_2 = 1{,}500$ $\mu\mu$f, demonstrated its working efficiency. The transformer Tr was wound on type Sh-12 laminations made of É4AA grade steel, with 500 turns in the primary winding w_1, and 3,000 turns in the secondary winding w_2 of 0.1 mm type PÉL wire. Capacitor $C_4 = 1\,\mu$f. The detector was of the DGTs-2 type. The polarized relay PR was of the RP-7 type with an operating current of 30 μ a. Resistance $R_a = 1.5$ kohm was of the wire-wound type and could be adjusted within $\pm 20\%$ limits.

The circuit used for sorting capacitors (Fig. 2) consists of a two-stage amplifier with a large negative feedback for stabilizing its gain and a twin RC-network, one-half of which is used for inserting the tested capacitors. By means of resistor R_1 the gain of the amplifier is adjusted in such a manner that for $C_{ts} < C_{rf}$ the circuit starts oscillating, but for $C_{ts} > C_{rf}$ the circuit stops oscillating. In other particulars the circuit is the same as the one already described.

In the above circuit it is advisable to use reference capacitors of the same type as the tested capacitors, but subjected to artificial ageing in order to stabilize their capacitance, thus excluding the effect of the ambient temperature on the accuracy of measurement. By providing a different positive feedback circuit for each

nominal value of tested capacity, we obtain similar type circuits, which is convenient both from the point of view of standardized units and a quick change for testing different types of capacitors.

The selection of resistors R and R_2, which are normally taken to be equal, is determined by the frequency of the test oscillation. For instance, when sorting out capacitors of a value smaller than 100μμf, the frequency of test oscillations f_0 should be in the range of 8 to 10 kc, and the value of resistor $R = R_2$ is obtained from the formula

$$R = \frac{1}{3.14 f_0 C_{rf}} \qquad (2)$$

When capacitors with a value between 100 and 1,000 μμf are being sorted out, the frequency of test oscillations should be in the range of 4 to 5 kc, and it should be further decreased with a rising value of testing capacitors, so as to keep their impedance calculated from (2) at not less than 500 ohm, since for lower values the output resistance begins to affect the measurements.

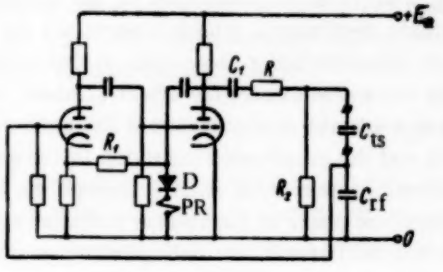

Fig. 2.

For testing inductances the circuit remains the same; only the positive feedback circuit is changed, namely, capacitance C_1 is substituted by an inductance equal to $2L_{rf}$, and C_{rf} and C_{ts} are substituted by L_{rf} and L_{ts} respectively. Moreover, the positive feedback LR-network is connected to the output of the amplifier through an isolating capacitor with a voltage drop across it small enough to be neglected as compared with the voltage drop across the elements of the LR-network. The value of resistor R should be calculated from the formula given below, but should not be less than 500 ohm:

$$R = 12.56 f_0 L_{rf}. \qquad (3)$$

For sorting coils with ferromagnetic cores which normally operate with superimposed dc magnetization, the dc source should be connected in series with the L_{rf} and L_{ts} circuits.

The relative simplicity of the above circuit makes it possible to use it for sorting resistances by their values. Basically the circuit remains the same, but resistor R_2 is used as a divider.

In practice the circuit consisted of a double triode amplifier type 6N2P, with $R = R_2 = 330$ kohm, $R_1 = 220$ ohm, detector D type DGTs-4 and polarized relay PR type RP-7 with an operating current of 30 μa.

Exhaustive tests have shown that the circuit operates satisfactorily in sorting out capacitors by their capacity, and coils by the value of their inductance or resistance.

Another noteworthy point consists of the small effect produced by supply voltage variations on the accuracy of sorting (less than 0.1% for 1% change in the supply voltage), which was achieved owing to the very large negative feedback (without the negative feedback the amplifier has a gain of the order of 1,000, and with it only 6). This provides a high operating reliability of the circuit, as well as better operating conditions for the electronic tubes comprising the circuit, thus increasing their life.

For testing components which require resistor R_2 to be less than 500 ohm it is necessary to use a circuit with cathode followers.

In the above circuits the stability of sorting amounts to 1-1.5%, which is normally satisfactory for mass-produced electronic tubes, capacitors, coils, and resistors.

For precise measurements of the slope of the anode characteristic in amplifying tubes a balanced circuit is used. In this circuit a variable reference resistance (a resistance box) is connected through an isolating capacitor to the anode of the tube under test; the other end of this resistance and the control grid of the tube are connected to a voltage source of 1,000 cps. In this circuit a difference of two voltages will appear at the anode of the tested tube, and for a certain value of the anode resistance, which corresponds to the slope of the tube under test, we shall obtain a balanced condition, which can be observed on a tube voltmeter connected to the anode of the tube. In practice, in addition to amplitude balancing the phase must also be balanced, thus complicating the process and making it inconvenient for automation. By applying the principle previously used in designing sorting circuits, it is possible to combine in one unit the functions of an oscillator and a tube voltmeter amplifier.

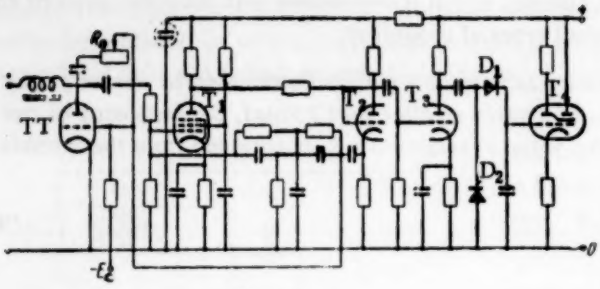

Fig. 3.

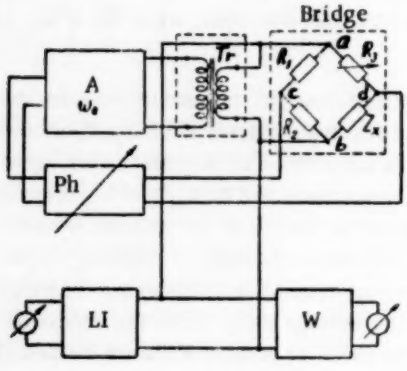

Fig. 4. A is an amplifier; Ph is a phase-shifting unit; LI is a level-indicator; W is a wave-meter.

Figure 3 shows the schematic of such a unit, which is an ordinary tuned amplifier. The input of the amplifier is connected to the anode of the tube under test. Its output is connected to a calibrated resistor in the anode circuit and to the control grid of the tested tube. In this circuit the tested tube acts in series with the positive feedback as an additional attenuation, whose transfer constant at the point of balance is equal to zero. It is obvious that if the attenuation in the positive feedback circuit is smaller than the amplifier gain, and the total phase shift in the loop consisting of the amplifier and the feedback circuit is equal to zero, sustained oscillations will arise in the circuit. The stopping of oscillations indicates an approach to the balanced condition. The circuit shown in Fig. 3 provides amplification for the reading of small amplitude oscillations, which is necessary for the type of slope characteristics measured. It should be noted that in the above circuit the effect of phase imbalance leads to a certain displacement in the generated oscillations and the consequent reduction in the gain of the amplifier. Moreover, in passing through the balanced condition the phase of the system is shifted by 180°, and hence oscillations are only possible on one side of the balance, thus providing a most favorable property.

Tests of the circuit shown in Fig. 3 for a gain of $K_{am} = 490$, $f_0 = 1000$ cps have revealed its satisfactory operational properties. The comparison of these results with the measurements which were obtained by means of a complete circuit containing an oscillator and an amplifier, and which were made without changing the operating conditions of the tube under test, showed a difference between the two methods of 0.2-0.3% which, for the given amplifier gain, is in good agreement with calculations (the value of the imbalance which starts oscillations is inversely proportional to the gain). It was also found that under the most adverse phase relations the decrease in the amplifier gain does not exceed 20%, which provides an additional error of 0.02%. It is, therefore, unnecessary to provide any phase compensation.

For more precise measurements of capacitances, inductances of coils and resistances it is advisable to use a bridge circuit.

A block-schematic of a bridge circuit with a unit which combines the functions of an amplifier and an oscillator is shown in Fig. 4. The principle of operation of this circuit is similar to the one already described, and consists of oscillations arising in the circuit when the bridge imbalance exceeds a certain limit, which is determined by the gain of the tuned amplifier. Thus, for an amplifier gain of 1,000 the value of this imbalance does not exceed 0.1%, which is permissible providing the error in the bridge arm resistance is of the order of 0.1%.

At present contactless methods are used for checking components by their conductance [3], which are based on measuring the parameters of tuned circuits with the measured components placed inside the coil of the circuit.

The measuring circuits used for this purpose consist of an oscillator, a bridge which contains in one of its arms the measuring coil, and an amplifier. The adjustment of these circuits presents a certain difficulty owing to the possible coupling between the oscillator and the amplifier, and to the necessity of compensating for unwanted phase shifts. Automatic measurements by means of this circuit provide even greater difficulties.

On the basis of the method described in this article we designed for checking the conductance of tungsten rods (variation in the conductance of these rods was caused by longitudinal cracks) a test circuit similar in its layout to the one shown in Fig. 4. We used a two-stage amplifier with electronic tubes type 6Zh1P and a gain

of the order of 1,000 in the range of 120-170 kc. Bridge components R_1 and R_2 consisted of 150 kohm resistors type VS-1.0, and elements Z_X and R_3 consisted of LC-circuits tuned to 150 kc. A sample whose conductance lies between the limits of "good" and "rejected" is placed during testing inside the coil of component Z_X, and the sample under test is placed in the coil of component R_3. Transformer Tr is air cored and has windings $w_1 = w_2 = 400$ turns, wound with 0.1 mm type PEL wire. The remaining elements are the same as those used in the circuits previously described. Tests carried out with the automatic circuit for determining the conductance of tungsten rods indicate the possibility of its practical application to other similar measurements.

It should be noted that when the variations of circuit parameters between good and rejected samples are small, it is necessary to use an amplifier with a larger gain or introduce an amplitude selector.

Engineers V. I. Krasavina, M. P. Koroleva, L. M. Shchukina, and L. N. Kusakin assisted in carrying out this work.

SUMMARY

The method described in this article for designing sorting circuits for electrical components provides the possibility of constructing simple and reliable automatic devices for sorting electronic tubes by the slope of their anode characteristics, capacitances; coils by their inductances and resistances; as well as various electrical components by their conductance.

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PHASE-SHIFTING BRIDGE WITH A POTENTIOMETER-CONTROLLED CONTINUOUS OPERATION

S. M. Katsnel'son

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The defects of many impedance phase-shifting bridges which serve to measure the variations of the phase of one voltage with respect to another consist in the limited phase-control range of the output voltage or the relative complexity of the bridge.

The author of this article proposes a new circuit for a phase-shifting bridge with a continuous control [1], attained by means of only one capacitor and one adjustable resistor used as a potentiometer.

By comparison with the existing designs the new phase-shifting bridge is simple and provides a continuous shifting of the output voltage phase in the range of 360° or 360° · n, where n is any arbitrary integer or fraction. The basic element of the circuit of a phase-shifting bridge which provides a smooth variation in the phase of the output voltage through an angle exactly equal to 180° is shown in Fig. 1. The secondary winding of a center-tapped transformer is connected to an adjustable resistor R_0 in the form of a potentiometer whose slide is connected through a fixed capacitor or inductor to one of the ends of the transformer winding.

The voltage vector between the mid-point of the transformer and the potentiometer slide can be represented by the equation

$$\bar{U}_V = \frac{\bar{U}_0}{1+\left(\frac{R_1R_2\omega C}{R_0}\right)^2}\left[\frac{1}{2}\left(\frac{R_1R_2\omega C}{R_0}\right)^2 + \frac{R_1}{R_0} - \frac{1}{2} + j\,\frac{R_1R_2\omega C}{R_0}\left(1-\frac{R_1}{R_0}\right)\right], \qquad (1)$$

*See English translation.

where $\overline{U}_0$ is the voltage across the secondary winding of the transformer.

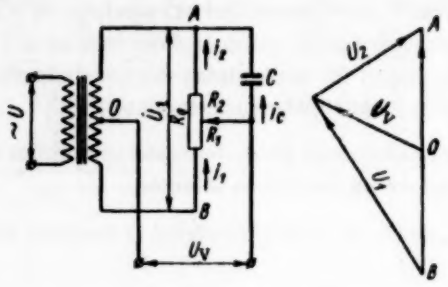

Fig. 1. Schematic of the impedance phase-shifting bridge with a resistance potentiometer and a vector diagram.

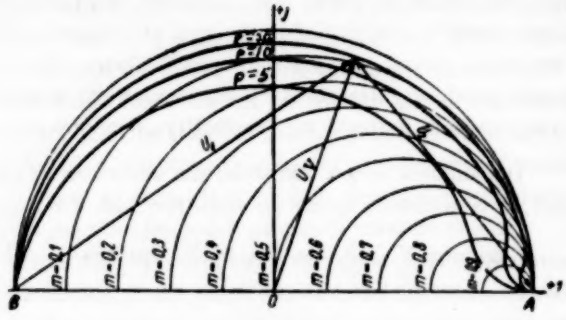

Fig. 2. Locus of the ends of the output voltage vector in a phase-shifting bridge with a potentiometer control.

When the potentiometer slide is displaced the end of vector $\overline{U}_V$ describes a curve which is distinct from a circle, but the angle of its rotation amounts to a complete 180° (Fig. 2). The deviation from a circumference of the curve described by the output voltage vector $\overline{U}_V$ depends on the ratio between the total resistance of potentiometer R_0 and the capacitive reactance $1/\omega C$. In this circuit the ratio of the total resistance to the capacitive reactance equals a constant p:

$$p = R_0 \omega C.$$

The family of curves shown in Fig. 2 represents the locus of the ends of the output voltage vector $\overline{U}_V$ for various values of p. For instance, for the value of p = 10, the variations of the output voltage vector amplitude do not exceed 10%. For smaller values of p variations in the amplitude of voltage $\overline{U}_V$ increase, but the irregularity in the relation between the rotation angle of the voltage vector and that of the potentiometer axis decreases. Denoting by m the ratio of R_1/R_0, let us find the locus of the ends of the output voltage vector for different values of p and a constant m. These curves are also shown in Fig. 2. They consist of a family of circles with a common point which corresponds to the end of the vector $\overline{U}_V$. In fact, by transferring the origin of the coordinates to point A and transforming (1) we obtain the value of vector $\overline{U}_2$:

$$\overline{U}_2 = \frac{\overline{U}_0 (1-m)}{1+m^2 (1-m)^2 p^2} [1 - jm (1-m) p]. \qquad (2)$$

The coordinates of the origin of this vector X and Y are respectively equal to

$$X = \frac{U_0 (1-m)}{1+m^2 (1-m)^2 p^2},$$

$$Y = \frac{U_0 (1-m^2) mp}{1+m^2 (1-m)^2 p^2}.$$

Then

$$X^2 + Y^2 = U_0 (1-m) X.$$

The above equality represents an equation of a circle of the following form:

$$(X-A)^2 + Y^2 = A^2,$$

where A is the radius of the circle, equal to the coordinate of its center along the X axis:

$$A = \frac{U_0 (1-m)}{2}.$$

Hence, for various values of m we obtain a family of circles whose center lies on the X axis so that the coordinate of the center is equal to the radius of a circumscribed circle. The tangent of the angle formed by vector $\overline{U}_2$ with the X axis is determined by the equation

$$\tan \psi_2 = (1-m)\, mp. \tag{3}$$

Then the locus of the ends of the output voltage vector can be easily determined graphically by plotting it as a function of m for a given value of $p = R_0 \omega C$.

The fact that the circuit of the phase-shifting bridge under consideration provides a phase shift in the output voltage vector $\overline{U}_V$ through an angle of exactly 180° makes it possible to use this circuit for continuous phase-shifting.

In fact, if the potentiometer slide is in one of its extreme positions, the switching of capacitance C from one end of the transformer winding to the other will not produce any change in the phase of the output voltage. However, when the slide is moved in the opposite direction we will obtain a rotation of vector $\overline{U}_V$ in the second half of the circle, thus completing the range of the controlled angle to 360°. By designing the construction of the phase-shifter in such a way that in the extreme positions of the potentiometer slide the capacitance C is switched from one end of the transformer winding to the other, we shall obtain a continuous phase-shifter rotating the phase of the output voltage through an angle of $360° \cdot n$, where n is any arbitrary number.

The construction of a continuous phase-shifter is shown in Fig. 3. The phase-shifter consists of a potentiometer 1, capacitor 2 and switch 3. The switch is arranged in such a manner that in the extreme positions of slide 4 the capacitor C is switched from one end of the transformer winding to the other. For this purpose an insulated tip 5 is connected to the end of the potentiometer slide and made to operate the switch at the instant when the slide reaches one of its extreme positions. By moving the potentiometer slide from one of its extreme positions to the other we obtain an output voltage phase rotation through an angle of $360° \cdot n$, where n is any arbitrary number. When the potentiometer slide passes from one extreme position to the other the vector continues to rotate in the same direction.

Let us now give an abridged analysis of the effect of loading and the stray reactance of the transformer winding on the output voltage phase.

The value of the output voltage $\overline{U}_V$ of the phase-shifting bridge, without taking into consideration the loading and stray transformer reactance, is easily determined by the nodal voltage method, according to which it is equal to

$$\overline{U}_V = \frac{\Sigma \overline{E}\, \overline{Y}}{\Sigma \overline{Y}}.$$

In our specific case it will be

$$\overline{U}_V = \frac{\frac{\overline{E}_0}{2}(g_2 + j\omega C - g_1)}{g_1 + g_2 + j\omega C}. \tag{4}$$

Here g_1 and g_2 are the conductances of the bridge arms equal to

$$g_1 = \frac{1}{R_1};\ g_2 = \frac{1}{R_2};$$

$\overline{E}_0$ is the emf of the transformer's secondary winding.

With loading, Eq. (4) takes the form

$$\overline{U}_{V1} = \frac{\frac{\overline{E}_0}{2}(g_2 + j\omega C - g_1)}{g_1 + g_2 + j\omega C + \overline{Y}_L}, \tag{5}$$

where $\overline{Y}_L$ is the admittance of the load.

The ratio of the output voltage with loading $\overline{U}_{V1}$ to the voltage on open circuit $\overline{U}_V$ is equal to

$$\frac{\overline{U}_{V1}}{\overline{U}_V} = \frac{1}{1+\dfrac{\overline{Y}_L}{g_1+g_2+j\omega C}} = \overline{K}. \tag{6}$$

The parameters of the circuit are determined under concrete conditions for a given load and a permissible decrease K in the output voltage during the setting of the phase angle.

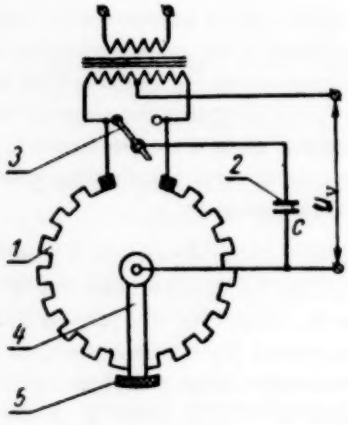

Fig. 3.

With a resistive load, ratio $\overline{U}_{V1}/\overline{U}_V$ is easily determined for the case when the sum of the adjustable admittances in the bridge attains a minimum:

$$g_1+g_2=\frac{1}{R_1}+\frac{1}{R_0-R_1},$$

and when

$$R_1=\frac{R_0}{2}; \quad g_1+g_2=\frac{4}{R_0}.$$

By inserting the value of the sum of the adjustable admittances thus obtained into (6) we have

$$\frac{\overline{U}_{V1}}{\overline{U}_V} = \frac{1}{1+\dfrac{g_L R_0}{4+j\omega C R_0}} = \overline{K}. \tag{7}$$

Let us now find the modulus of the denominator in (7):

$$\left|1+\frac{g_L R_0}{4+j\omega C R_0}\right| = \sqrt{\frac{(g_L R_0)^2+8g_L R_0+16+p^2}{16+p^2}},$$

where $p = \omega C R_0$.

Then, by solving the quadratic equation

$$(g_L R_0)^2+8g_L R_0+16+p^2=\frac{16+p^2}{k^2},$$

where $|k|$ is the modulus of the ratio $\overline{U}_{V1}/\overline{U}_V$, we obtain an equation relating the parameters of the bridge with the value of the load and the permissible reduction in the voltage:

$$g_L R_0=\frac{R_0}{R_L}=-4+\frac{1}{k}\sqrt{16+p^2(1-k^2)}. \tag{8}$$

It should, however, be taken into account that on open circuit the minimum output voltage is different from the voltage $U_0/2$ at the secondary transformer winding as shown in Figs. 2 and 4. Hence, by expressing all the voltages in relative units and referring them to the voltage $U_0/2$ at the secondary winding of the transformer, we shall obtain the minimum relative open circuit voltage U^*_{V0} from Fig. 4 or for $p \geq 5$ from the approximate formula

$$U^*_{V0} = 1 - \frac{p}{p^2 - 1}.$$

The minimum relative output voltage with a load will be

$$U^*_{VL} = k'kU^*_{V0} = k'k\left(1 - \frac{p}{p^2 - 1}\right), \tag{8a}$$

where k is a coefficient included in (8) and accounting for the lowering of the voltage by the effect of loading; and k' is the correction coefficient (0.98-1.0).

In calculating the bridge parameters quantity p should be selected in the range of 10-20, since for smaller values of p a large decrease in the open circuit voltage is obtained.

The bridge can be calculated on the basis of various basic data. For instance, by taking as the basis a load resistance R_L, a decreasing-voltage coefficient k accounting for the transition from open circuit to operating conditions under a load and choosing quantity p = 10-20, it is easy to find the parameters of the bridge and determine the minimum relative voltage under load U^*_{VL}.

Example. For R_L = 10 kohm, p = 10, k = 0.9 the bridge resistance R_0 is determined from (8):

$$R_0 = 10\left[-4 + \frac{1}{0.9}\sqrt{16 + 100(1 - 0.9^2)}\right] = 25.8 \text{ kilohm}.$$

Capacitance C will be

$$C = \frac{p}{R_0\omega} = \frac{10}{25.8 \cdot 10^3 \cdot 314} = 1.23 \ \mu\text{f}.$$

The minimum relative voltage is found from (8a):

$$U^*_{VL} = k'k\left(1 - \frac{p}{p^2 - 1}\right) = 0.98 \cdot 0.9\left(1 - \frac{10}{99}\right) = 0.795.$$

The difference between the output voltage phase under load and that of the open circuit voltage in a general case is determined from (6). With a resistive load the phase difference between these voltages is

$$\psi_L = \tan^{-1} \frac{g_L\omega C}{(g_1 + g_2 + g_L)(g_2 + g_1) + (\omega C)^2}. \tag{9}$$

The positive sign of ψ_L shows that with a resistive load the output voltage vector $\overline{U}_{V1}$ will lead the open circuit voltage vector $\overline{U}_V$ by an angle of ψ_L. The phase characteristic of the bridge is then improved, i.e., the relation between the output voltage phase-shift and the rotation angle of the potentiometer axis becomes more linear. The maximum value of angle ψ_L is easily found by differentiating the denominator in (9) with respect to R_1 and equating the first derivative to zero. We shall then obtain $R_1 = R_0/2$. Hence

$$\psi_{L\,max} = \tan^{-1} \frac{\omega C}{R_L\left(\frac{16}{R_0^2} + \frac{4}{R_0 R_L} + \omega^2 C^2\right)}.$$

Let us now examine the effect of the transformer winding's stray reactance x_s on the output voltage phase. The output voltage without a load, but taking into account the transformer winding's stray reactance will be

$$\bar{U}_{V2} = \frac{\frac{E_0}{2}\left(\frac{1}{j\frac{x_s}{2} + \frac{1}{g_2 + j\omega C}} - \frac{1}{\frac{1}{g_1} + j\frac{x_s}{2}}\right)}{\frac{1}{j\frac{x_s}{2} + \frac{1}{g_2 + j\omega C}} + \frac{1}{\frac{1}{g_1} + j\frac{x_s}{2}}}. \tag{10}$$

After transformation this equation takes the form

$$\bar{U}_{V2} = \frac{\frac{\bar{E}_0}{2}(g_2 - g_1 + j\omega C)}{g_1 + g_2 + j\omega C + jx_s g_1 (g_2 + j\omega C)}. \tag{11}$$

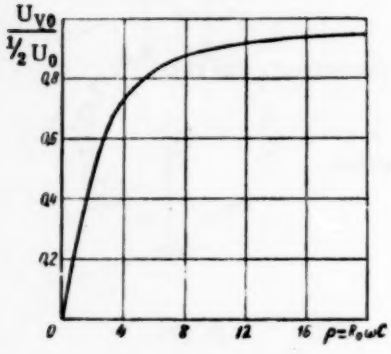

Fig. 4. Graph showing the relation between the bridge open-circuit voltage $\frac{U_{V0}}{\frac{1}{2}U_0}$ and the bridge parameters $p = R_0\omega C$.

Taking the ratio of the open circuit voltage from (4) to the voltage determined from (11), we find

$$\frac{\bar{U}_V}{\bar{U}_{V2}} = 1 + \frac{jx_s g_1}{1 + \frac{g_1}{g_2 + j\omega C}}.$$

For the two extreme positions of the adjustable resistance R_0 we shall find that in one case $g_1 = g_0$ and $g_2 = \infty$. Then

$$\frac{\bar{U}_V}{\bar{U}_{V2}} = 1 + jx_s g_0.$$

In the second case for $g_2 = g_0$ and $g_1 = \infty$ we have

$$\frac{\bar{U}_V}{\bar{U}_{V2}} = 1 + jx_s(g_0 + j\omega C) = 1 - x_s\omega C + jx_s g_0.$$

An analysis of these expressions shows that the output voltage phase changes due to the effect of the transformer winding's stray reactance are partly compensated in both extreme positions of the adjustable resistor. In fact, the output voltage vector phase shift, taking into account the transformer winding's stray reactance, is determined with respect to the emf of the secondary winding in one of the extreme positions of the adjustable resistor by angle ψ_{s1} which is equal to

$$\psi_{s1} = \tan^{-1} x_s g_0,$$

and in the other extreme position by angle ψ_{s2}:

$$\psi_{s2} = \tan^{-1} \frac{x_s g_0}{1 - x_s\omega C}.$$

This phase shift in both extreme positions of the adjustable resistor is in the same direction; hence the difference of the total rotation angle of the output voltage vector from 180° is determined by angle ψ_{ss}:

$$\psi_{ss} = \tan^{-1} \frac{x_s g_0}{1 - x_s\omega C} - \tan^{-1} x_s g_0. \tag{12}$$

This angle is very small so that in most practical cases it can be neglected.

SUMMARY

Models of the above continuously controlled phase-shifting bridges with one reactance and a resistive potentiometer were used in grid control circuits of gas rectifiers, in synchronizing circuits of cathode ray oscilloscopes and in other instruments. The output voltage phase displacement is very smooth, without jumps at the extreme position of the potentiometer slide. The exceptional simplicity of the continuously controlled phase-shifting bridge with a resistive potentiometer, as compared with other known types which provide a continuous output voltage phase shift through an angle of $360° \cdot n$ [2], will lead to its wide application in control circuits of electronic and gas rectifiers, as well as in experimental and laboratory practice.

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UTILIZATION OF OPERATIONAL AMPLIFIERS FOR RAISING THE INPUT IMPEDANCE OF MEASURING CIRCUITS

L. I. Dyatlov and G. M. Petrov

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In measurements as well as other spheres, for instance, in electronic simulation, an error of considerable importance is often produced by connecting a measuring circuit with an insufficiently high input impedance to a low-power source of voltage under test. In particular, this problem arises when it is required to measure with an error not exceeding $\pm 0.1\%$ the voltage across the sliders of a multiplying potentiometer with a resistance of the order of 40-100 kohm by means of an electronic digital voltmeter with an input impedance of the order of 1 meg.

It will be easily seen that if the electronic digital voltmeter is connected directly to the multiplying potentiometer the error due to loading will considerably exceed the specified limits.

In order to avoid the error of loading it is necessary to have a linear element, which we shall henceforth call a matching circuit, with a transfer constant of $1 \pm (0.01\text{-}0.1)\%$, an input impedance not less than 100 meg and a small zero drift (for instance, not exceeding 1 mv). For the above purpose two special amplifier circuits were designed on the principle of using modulation, amplification and demodulation, with two parallel channels for signals of a high and low frequency [1].

It is shown in this article that the matching circuit can be made up of two normal operational dc amplifiers with a positive feedback. The main advantage of this circuit consists in using standard operational dc amplifiers, which are normally used in electronic simulating devices.

Below we show the schematic of the matching circuit (Fig. 1), derive its basic equations and provide an analysis of errors due to the inaccuracy in the manufacture of resistors and a zero drift and grid current in the operational amplifiers.

From the theory of operational dc amplifiers with a large negative amplification factor [2, 3] it is known that, with a large negative feedback, the relation between the input and output voltages in these amplifiers is determined only by the parameters of the feedback circuit and does not depend on the parameters of the amplifying channel. Moreover, owing to the negative feedback the output impedance of the circuit is reduced to a very small value, in practice smaller than 1 ohm.

In the circuit shown in Fig. 1 there are two such amplifiers, A_1 and A_2. The basic equation which represents the circuit in Fig. 1, without taking into consideration the drift emf and the amplifier grid currents, has the following form:

$$U_1 = -\frac{R_2}{R_1} U'_{in}; \quad i_{in} = i_1 - i_2;$$

$$U_2 = -\frac{R_4}{R_3} U_1; \quad i_1 \simeq \frac{U'_{in}}{R_1}; \qquad (1)$$

$$R_{in} = \frac{U'_{in}}{i_{in}}; \quad i_2 = \frac{U_2 - U'_{in}}{R_5}.$$

If we introduce the notations $R_2/R_1 = K_1$ and $R_4/R_3 = K_2$, the equation for the input impedance of the matching circuit can be represented after simple transformations in the form

$$R_{in} = \frac{U'_{in}}{i_{in}} = \frac{R_5}{\frac{R_5}{R_1} - (K_1 K_2 - 1)}. \qquad (2)$$

It will be seen from (2) that for $R_5/R_1 = (K_1K_2-1)$ the input impedance of the circuit will be infinitely large, i.e., the circuit shown in Fig. 1 will in principle provide ideal matching.

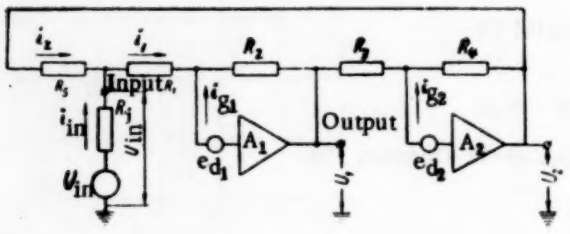

Fig. 1. A_1 and A_2 are operational dc amplifiers; R_1, R_2, R_3, R_4 and R_5 are feedback resistances; e_{d1} and e_{d2} are drift emfs referred to the amplifier inputs; i_{g1} and i_{g2} are the amplifier grid currents; U_{in} is the input voltage; U_1 is the output voltage of the matching circuit; R_i is the internal impedance of the input voltage source.

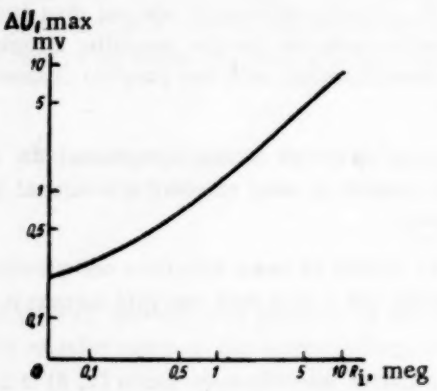

Fig. 3.

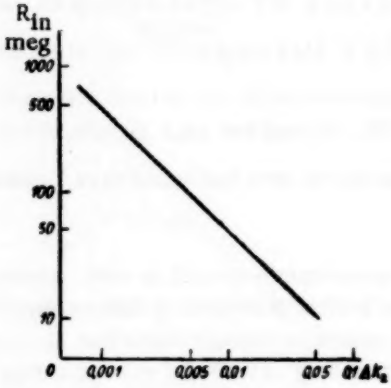

Fig. 2.

However, due to the limited accuracy in making the resistors used in feedback circuits, various quantities, such as K_2, will always differ by a certain amount ΔK_2 from their ideal value

$$K_{2\infty} = \left(\frac{R_5}{R_1} + 1\right)\frac{1}{K_1},$$

for which the input impedance is infinite.* The expression for the input impedance will then take the form

$$R_{in} = \frac{R_5}{\Delta K_2} \cdot \frac{1}{K_1}. \qquad (3)$$

The case when the amplification factor K_1 of the first amplifier is equal to 1 is of practical interest. In this condition the circuit can be used as a precision inverter with a large internal impedance.

*Here for simplicity only the inaccuracy of coefficient K_2 is examined. The value of ΔK_2 should include a component due to the possible inaccuracy of coefficient K_1. Then it is possible to consider (for simplicity of reasoning) that coefficient K_1 is ideally accurate.

Figure 2 shows a graph of the relation between the circuit input impedance and quantity ΔK_2 for $K_1 = 1$ and $R_5 = 0.5$ meg, from which it will be seen that for $\Delta K_2 \leq 0.005$ the input impedance $R_{in} \geq 100$ meg. In practice there is no difficulty in providing the value of the operational coefficient K_2 with the required accuracy by means of precision wire-wound resistors.

Expressions (2) and (3), which hold for the case when the amplifier has no zero drift or grid current, show that with a decreasing value of $|\Delta K_2|$ the matching circuit input impedance rises and in the limiting case for $|\Delta K_2| \to 0$, impedance $R_b \to \infty$.

In practice it is, however, necessary to account for the sources of interference always present in operational amplifiers and normally referred for calculation purposes to an emf and current operating in the input circuit.

In the presence of a drift emf (referred to the input of the amplifier as shown in the circuit) and a grid current in the amplifier the circuit equations can be written with a sufficient degree of accuracy in the following approximate form:

$$U_2 = -U_1 \frac{R_4}{R_3} - e_{d2}\left(1 + \frac{R_4}{R_3}\right) - i_{g2}R_4;$$

$$U_1 = -U'_{in}\frac{R_2}{R_1} - e_{d1}\left(1 + \frac{R_2}{R_1}\right) - i_{g1}R_2;$$

$$U'_{in} = \frac{(U_{in}R_5 + U_2R_i)R_1}{R_1R_5 + R_iR_5 + R_iR_1}. \tag{4}$$

After inserting the previously adopted notations for R_2/R_1 and R_4/R_3 and certain simple transformations, the expressions for the output voltage U_1 will take the form

$$U_1 = \frac{-U_{in}R_5R_2 + [e_{d2}(1+K_2) + i_{g2}R_4]R_iR_2}{R_1R + R_iR_5 + R_iR_1 - K_2R_iR_2} - \frac{R_1R_5 + R_iR_5 + R_iR_1}{R_1R_5 + R_iR_5 + R_iR_1 - K_2R_iR_2} \times$$

$$\times [e_{d1}(1+K_1) + i_{g1}R_2]. \tag{5}$$

If the internal resistance R_i of the input signal source and the input voltage U_{in} are equal to zero, the error in the circuit will only depend on the grid current and the zero drift of the amplifier A_1:

$$\Delta U_1 = -e_{d1}(1+K_1) - i_{g1}R_2. \tag{6}$$

If the finite internal resistance of the source and the deviation of coefficient K_2 from its ideal value by ΔK_2 are taken into account, formula (6) takes the form

$$\Delta U_1 = \frac{e_{d2}(1+K_2) + i_{g2}R_4}{\beta + \Delta K_2} - \frac{\beta + K_2}{\beta + \Delta K_2}[e_{d1}(1+K_1) + i_{g1}R_2], \tag{7}$$

where

$$\beta = \frac{R_5}{R_i}.$$

It will be seen from (7) that the circuit error depends both on the value of ΔK_2 and the ratio of resistances R_5 and R_i.

Deviation ΔK_2 has a pronounced effect on the circuit input impedance, and it should be made therefore as small as possible. If the value of ΔK_2 is made considerably smaller than β the circuit error due to the drift emf and the grid current will be in the main determined by the ratio of resistances R_5 and R_i.

The fact that the internal resistance R_i affects the error ΔU_1 even when the input impedance of the circuit is equal to infinity ($\Delta K_2 = 0$) is explained by the following considerations.

In an ideal case, i.e., when there is no interference in the circuit and $\Delta K_2 = 0$, then $R_{in} = \infty$ and $U'_{in} = U_{in}$. The noise voltage which arises anywhere in the circuit and causes an error in the output voltage U_1 is fed to the input of the amplifier A_1 over the feedback circuit and through the potential divider formed by resistances R_s and R_i. The greater the value of R_i the larger will be the portion of the noise voltage which is fed to the input of amplifier A_1 and forms part of the output circuit voltage U_1.

Figure 3 shows the relation between the absolute error of the amplifier and the internal resistance R_i of the signal source for $R_s = 0.5$ meg, $e_{d1} = e_{d2} = 50$ μv, $i_{g1} = i_{g2} = 10^{-10}$ amp, and $|\Delta K_2| \ll |\beta|$.

It will be seen from Fig. 3 that for $R_i \leq 1$ meg the circuit error in this case will be $\Delta U_1 \leq 1.5$mv (0.0015% of 100 v). The input impedance of the circuit will then be practically infinite, since ΔK_2 can be made extremely small.

Thus, the internal impedance of the input signal source does not affect the internal impedance of the matching circuit, but affects considerably the error due to the zero drift and the grid current of the amplifier.

It is, therefore, advisable to use matching circuit amplifiers with as small a zero drift and grid current as possible. Of the dc amplifiers the most suitable for this purpose is amplifier type UPT-15 [4] developed by the electrical simulation department of the Scientific Research Institute of Computers or the amplifier with the separating capacitance at its input and a small grid current [5].

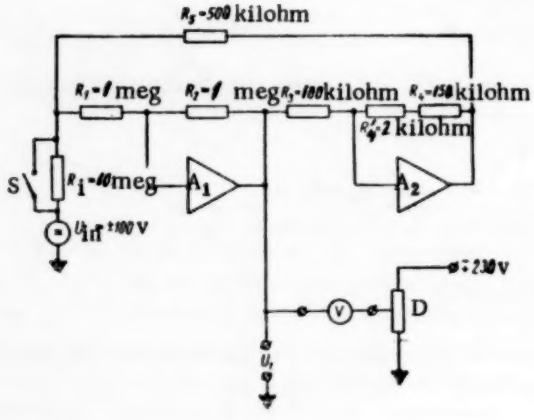

Fig. 4.

In order to check the above theoretical conclusions the circuit was tested out in the Scientific Research Institute of Computers (Fig. 4).

The tests included determining the relation between the circuit input impedance and ΔK_2, and the circuit error due to R_i. For determining the input resistance a dc voltage of 100 v was fed through a resistor of 10 meg to the input of the circuit. The voltage at the amplifier output was measured by a balancing method. After balancing out completely the voltage at the output of the amplifier, the value of the internal resistance was changed in steps from 10 meg to 0. The input resistance of the circuit was determined from the output voltage variations ΔU:

$$R_{in} = \frac{U_{in} - \Delta U}{\Delta U} R_i .$$

The value of ΔK_2 was varied by means of a wire-wound potentiometer $R_4 = 2$ kohms. In practice it was found easy to adjust the circuit parameters so as to make the input resistance not less than 1000 meg for an operation coefficient error not exceeding ±0.1% and a zero drift at the output not greater than ±0.005 v. The results thus obtained are in good agreement with theoretical conclusions.

SUMMARY

The analysis and testing of a matching circuit consisting of two operational dc amplifiers have shown that it is possible by means of this circuit to match dc voltage circuits with a large internal resistance (up to 10 meg) to circuits with a small internal resistance (up to 10 kohm).

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HIGH AND ULTRAHIGH FREQUENCY MEASUREMENTS

ELIMINATING ERRORS IN Q-METER TYPE KV-1

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It is usual to take the Q factor of a tuned circuit to be the ratio of the energy stored in the circuit to that lost over one period. For an ideal tuned circuit, which consists of three lumped elements: capacitance C, inductance L and resistance R, the Q factor is determined by the expression

$$Q=\frac{\omega L}{R}=\frac{1}{\omega CR}. \tag{1}$$

In certain cases it is more convenient to consider the resistance as the sum of two components, the resistance of the induction coil R_L and that of the capacitor R_C, both of which are equivalent to the losses in these elements; then (1) is slightly changed and assumes the form

$$\frac{1}{Q}=\frac{1}{Q_L}+\frac{1}{Q_C}.$$

where Q_L and Q_C are the Q factors of the induction coil and the capacitor.

Under actual conditions only the capacitor approaches fairly closely to the ideal condition and then only at relatively low frequencies.

The induction coil always has a residual interwinding capacitance C_0 and considerable losses, represented by a fairly complicated equivalent circuit (Fig. 1a).

For a low quality insulation material of the wire and the former, resistance R_0, which is equivalent to the losses in the insulation, can be comparable to resistance R_L.

In normal practice a further simplification is made by assuming all the losses to be concentrated in R_L. The equivalent circuit is then further simplified and assumes the form shown in Fig. 1b.

The Q factor of such a coil can be calculated from (1), but for this purpose it is necessary to determine all the coil elements, which is rather difficult. It is much simpler to assign to the coil an effective Q factor and represent it as shown in Fig. 1c, consisting of an impedance with an imaginary part $X=\omega L'=\frac{\omega L}{1-\omega^2 LC_0}$, and a real part $R'=\frac{R_L}{(1-\omega^2 LC_0)^2}$. These components provide the effective Q factor of the coil:

$$Q_{\text{eff}}=\frac{X}{R'}, \tag{2}$$

or

$$Q_{\text{eff}}=Q(1-\omega^2 LC_0). \tag{3}$$

Formula (2) is more convenient, since the real and the imaginary parts of the coil impedance can be easily determined by the resonance method using a Q meter. This instrument under certain conditions can measure the Q factor directly.

For determining the parameters of a coil it is connected to the measuring capacitor of the Q meter (Fig. 2), and voltage $\underline{e}$ is then impressed on the tuned circuit thus obtained. At resonance the following condition holds:

$$X=\omega L_x' = \frac{1}{\omega C_k}, \tag{4}$$

and the voltage U_k attains its maximum value equal to

$$U_k = \frac{e}{R_x'} \cdot \frac{1}{\omega C_k} = e\frac{X}{R_x'}. \tag{5}$$

From (4) and (5) it is easy to obtain the values of X and R_x and their ratio

$$\frac{U_k}{e} = \frac{X}{R_x'}. \tag{6}$$

This ratio is known as the voltage multiplication factor of a tuned circuit and it represents the ratio of the voltage magnitude across the condenser to that introduced into the circuit.

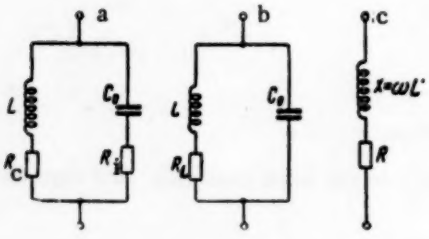

Fig. 1. R_C is the resistance equivalent to the copper losses in the coils; R_i is the resistance equivalent to the insulation losses; L is the inductance of the coil; C_0 is the interwinding capacitance of the coil.

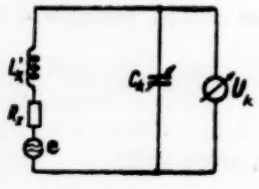

Fig. 2. Idealized schematic of the measuring circuit of instrument KV-1.

Sometimes the term "voltage multiplication," which characterizes the whole tuned circuit, is incorrectly used with respect to the coil alone, because it is represented, as shown in (6), by the parameters of the coil. It is more correct to describe the properties of a coil alone by means of the Q factor, and reserve the term "voltage multiplication" for a circuit which has in addition to a coil a capacitor and facilities for introducing and measuring voltages whose ratio is represented by the voltage-multiplication factor read off a pointer instrument of the Q meter in the form of a Q_n reading, which is also nominally known as the Q factor.

The differences between the voltage-multiplication factor and the effective Q factor are due to the effect of additional elements in the Q meter measuring circuit. Their value at low frequencies is relatively small compared with the parameters of the induction coil, and in this case the actual Q factor Q, the effective Q factor Q_{eff} of the coil, and the voltage-multiplication factor Q_n of the Q meter are sufficiently close to each other.

At high frequencies the induction coil parameters become of the same order of magnitude as the additional elements of the circuit, and the conversion from one of these quantities to the other requires a series of corrections.

In order to find these corrections let us examine the equivalent tuned circuit (Fig. 3) of the most commonly used Q meter type KV-1.

The voltage is introduced into the measuring circuit of the KV-1 meter by means of a coupling circuit consisting of resistor r_c, across which voltage $\underline{e}$ is generated by current I_n.

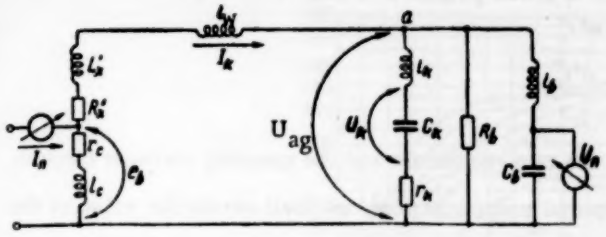

Fig. 3. L'_x and R'_x are the effective inductance and resistance of the measured coil; l_c and r_c are the residual inductance and resistance of the coupling element; C_k, l_k, and r_k are the capacitance of the measuring capacitor and its residual inductance and resistance; C_b, l_b and R_b are the capacitance, inductance and resistance of the tube-voltmeter input circuit "reading Q_n"; l_w is the residual inductance of the wiring and connections; I_n is the supply current read on the thermal ammeter "multiplier Q_n."

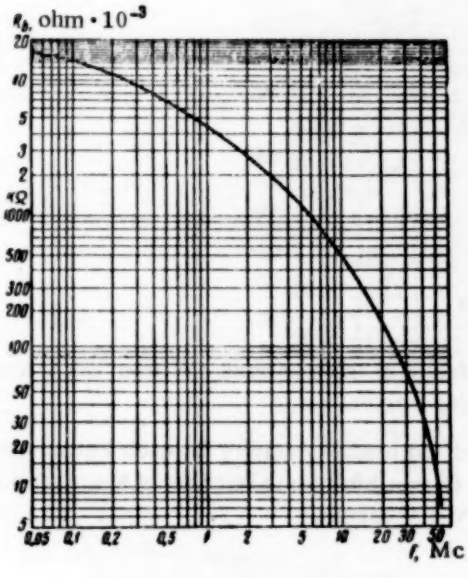

Fig. 4. Averaged relationship $R_b = \psi(f)$ for Q meters type KV-1.

Providing $r_c \ll R'_x + r_k + r_b$ and $\omega^2 l_c^2 \ll r_c^2$ the voltage thus inserted will be equal at low and medium frequencies to $e_i = I_n r_c$ and at high frequencies:

$$e_b = e_i \left(1 + \frac{\omega^2 l_c^2}{2r_c^2}\right). \qquad (7)$$

The value of e_i is set when the instrument is calibrated and corresponds to the reading of the meter "multiplier Q_n" equal to unity.

When the circuit is tuned to resonance the current through it will be determined by expression

$$I_k = \frac{e_b}{R_x + r_c + r_b + r_k}, \qquad (8)$$

where $r_b = \dfrac{1}{\omega^2 C_k^2 R_b}$ is the resistance introduced into the tuned circuit by the effect of the shunting resistor R_b.

If the reactive components are made to fulfill the condition that

$$\omega L'_x + \omega (l_k + l_w + l_c) - \frac{1}{\omega C_k} = 0, \qquad (9)$$

the voltage across the capacitor, taking into consideration its inductance, will be determined by

$$\bar{U}_{ag} = e_b \frac{1 - \omega^2 l_k C_k}{j\omega C_k (R_x + r_k + r_b + r_c)}. \qquad (10)$$

This expression has been obtained with certain simplifications which do not introduce noticeable errors: the impedance of the capacitor is represented by its reactance only, since its Q factor is always at least 1,000; the effect of the tube-voltmeter shunting the capacitor is not taken into account, since $C_k > C_b$.

After certain transformations of (10) making use of (9) we obtain

$$J\left|\frac{U_{ag}}{e_b}\right| = \frac{1 - \omega^2 C_k l_k}{\dfrac{1}{Q_{eff}}\left[1 - \omega^2 C_k (l_k + l_w + l_c)\right] + \omega C_k (r_k + r_b + r_c)} \qquad (11)$$

By expressing U_{ag} in terms of U_n – the indication of the tube-voltmeter "reading Q_n" – and ignoring the phase shift between e_b and U_n, we can write

$$\frac{U_{ag}}{e_b} = \frac{U_n}{e_i} \cdot \frac{1 - \omega^2 l_b C_b}{1 + \dfrac{\omega^2 l_c^2}{2r_c^2}}. \qquad (12)$$

Let us denote by $Q_n = \frac{U_n}{e_1}$ the Q factor value as read on the Q meter pointer instrument; and by

$$K = \frac{1-\omega^2 l_c^2}{1+\frac{\omega^2 l_c^2}{2r_c^2}}$$

the factor which corrects the effect of reactances in the tube-voltmeter and the coupling element circuits.

After substituting (12) in (11) and performing the required transformations, we shall obtain the value of the measured Q factor:

$$Q_{\text{eff}} = \frac{1-\frac{\omega^2 C_k (l_W+l_c)}{1-\omega^2 C_k l_k}}{\frac{1}{Q_n K} - (r_c+r_b+r_u)\frac{\omega C_k}{1-\omega^2 C_k l_k}}, \tag{13}$$

or

$$Q_{\text{eff}} = \frac{N}{\frac{1}{Q_n K} - \frac{1}{Q_k}}, \tag{14}$$

$$Q_{\text{eff}} = \frac{Q_n Q_k K N}{Q_k - Q_n K}, \tag{15}$$

where

$$N = 1 - \frac{\omega^2 C_k (l_W+l_c)}{1-\omega^2 C_k l_k};$$

and the Q factor of the Q meter itself:

$$Q_k = \frac{1-\omega^2 C_k' l_k}{(r_c+r_b+r_k)\omega C_k}.$$

Expressions (14) and (15) serve to calculate the effective Q factor of the measured coil from the Q meter readings taking into account the residual parameters of its measuring circuit.

The determination of these parameters is rather complicated and hardly necessary for each instrument; it is sufficient to use their averaged values, obtained by the NGIMIP (Novosibirsk State Institute of Measures and Measuring Instruments) from the test results of a whole series of KV-1 Q meters.

The parameters thus obtained were

$$l_W = 15 \cdot 10^{-9} \text{ h}; \quad l_k = 6.5 \cdot 10^{-9} \text{ h}; \quad l_c = 0.15 \cdot 10^{-9} \text{ h};$$
$$l_b = 180 \cdot 10^{-9} \text{ h}; \quad r_c = 0.04 \text{ ohm}; \quad r_k = 0.01 - 0.015 \text{ ohm};$$
$$C_b = 6 \ \mu\mu\text{f}.$$

The value for r_k given above refers to instruments produced since 1955; the KV-1 Q meters manufactured before that date have a larger r_k which varies from instrument to instrument.

The values of the tube-voltmeter input impedances for various frequencies are given in Fig. 4.

From the referred values of the residual parameters the relation of the set Q factor Q_k to C_k and frequency was calculated and plotted as shown in Figs. 5 and 6. The values of N and K are shown in Figs. 7 and 8.

It will be seen from Figs. 5 and 6 that the Q factor of the KV-1 Q meter measuring circuit is not large. Its highest value of 6600 units is obtained at about 500 kc with a circuit capacitance of 450 $\mu\mu$ f. At all the other frequencies and capacitances the Q factor is smaller and may even reach 120-150.

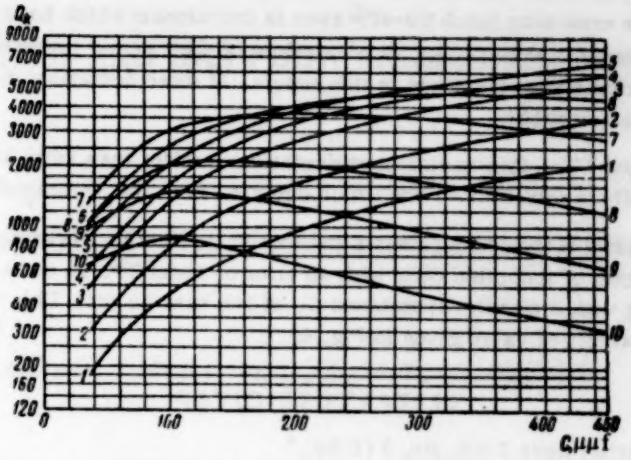

Fig. 5. Relation of Q factor Q_k to capacitance C_k for various frequencies f.

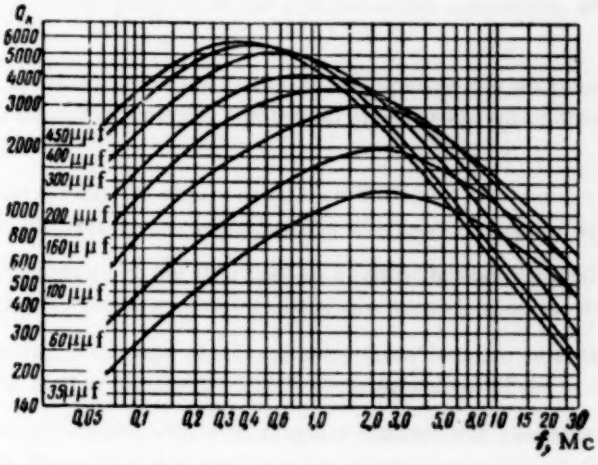

Fig. 6. Relation of Q factor Q_k to frequency f for various values of capacitance C_k.

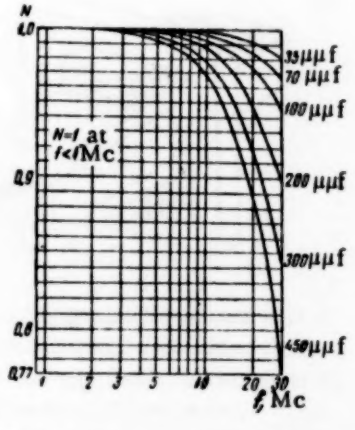

Fig. 7. Relation of factor N to frequency for various values of C_k.

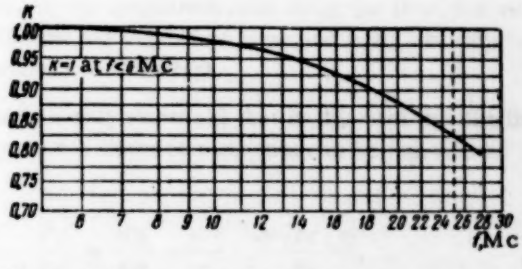

Fig. 8. Relation of the correcting factor K to frequency.

This feature of the KV-1 set should always be kept in mind and corrections introduced from (14) and (15). By means of these corrections it is possible to express the voltage-multiplication factor Q_n read on the meter in terms of the effective Q factor of the coil Q_{eff}.

The use of correction will lower the error in determining the actual Q factor by means of KV-1 Q meters to 10%. Without corrections the error may reach 20-40% even in instruments which have passed State testing. State testing of KV-1 Q meters consists in determining the deviations of their actual properties from those of an "averaged" instrument, for which correction curves are given in this article. If these deviations do not exceed 10% the instrument is considered to be satisfactory [1].

A large number of tested KV-1 instruments have deviations smaller than 10%. It is possible by means of a set of reference induction coils to select KV-1 sets which deviate from the "averaged" values by 1-2%.

By means of such specially selected instruments and by the application of corrections according to (15) it is possible to measure the Q factor of induction coils with an error of about 3%. In certain KV-1 models the error rose to 5-7% when measuring with a circuit capacitance C_k of less than 50 $\mu\mu$f. This is due to the deviation of the actual value of R_b from the averaged value given in Fig. 4.

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ELECTROMAGNETIC FLOWMETER WITH A RECTANGULAR CHANNEL

L. M. Korsunskii

Translated from Izmeritel'naya Tekhnika, No. 10, pp. 56-60, October, 1960

Electromagnetic flowmeters are based on the phenomenon of electromagnetic induction. Liquids flowing through a magnetic field induce in a transverse direction an emf which serves to measure the flow.

However, electromagnetic flowmeters do not measure the induced emf directly, but only the difference of potential u between the electrodes. This difference of potentials is affected by the circulating (equalizing) currents due to the presence of a potential gradient in the channel of the instrument, and also by various secondary factors.

The speed of the liquid is maximum in the center of the pipeline and decreases to zero at its sides. The velocity gradient in the radial direction produces a potential gradient and circulating currents in a plane perpendicular to the axis of the pipeline.

It was shown in [1, 2] that for axially symmetrical distribution of velocity the readings of an instrument with a circular channel and point electrodes do not depend on the distribution curve of velocities.

Any disturbances in the symmetry of the flow lead to a redistribution of circulating currents, which make the readings of the electromagnetic flowmeter with point electrodes dependent on the distribution of velocities [3, 4].

Under commercial operating conditions axial symmetry often does not exist in the pipes, and instruments with point electrodes cannot be used as flowmeters which are independent of the conditions of the flow.

In [5] an electromagnetic flowmeter with a rectangular channel and electrodes which equalize the electric field was suggested (Fig. 1). A theoretical and experimental investigation of such flowmeters was conducted in the KhGIMIP (Khar'kov State Institute of Measures and Measuring Instruments).

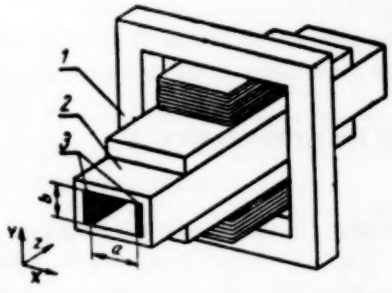

Fig. 1. Electromagnetic flowmeter with a rectangular channel and electrodes which equalize the electric field. 1) Magnet; 2) flowmeter chamber; 3) electrodes.

In the first approximation a stationary two-dimensional problem was examined (the magnetic field is stationary with time and uniform, all the conditions hold along the flow, the velocity of the liquid does not depend on time and is an arbitrary function of the transverse co-ordinates).

It can be easily shown [1] that in this case the distribution of potentials in the flowmeter channel can be represented by equation

$$\frac{\partial^2\varphi}{\partial x^2}+\frac{\partial^2\varphi}{\partial y^2}=\frac{\mu H}{c}\cdot\frac{\partial v_z(x,\,y)}{\partial x}, \qquad (1)$$

where φ is the potential; H is the magnetic field strength; μ is the permeability; c is the speed of light; v_z is the component of the velocity of the liquid along the channel of the instrument.

The cross section of the instrument channel is shown schematically in Fig. 2.

The boundary conditions in solving (1) are determined by the following considerations:

1. Near the walls of the instrument channel the liquid is stationary.

2. The walls of the channel, which are perpendicular to the magnetic field, are made of insulated material with a conductance $\gamma = 0$:

$$\frac{\partial \varphi}{\partial y}\Big|_{y=0} = \frac{\partial \varphi}{\partial y}\Big|_{y=b} = 0.$$

3. The walls of the channel which serve as electrodes are parallel to the magnetic field and are made from an "ideal" ($\gamma = \infty$) conductor:

$$\varphi|_{x=0} = \text{const} = \Phi_1;$$
$$\varphi|_{x=a} = \text{const} = \Phi_2.$$

However, these constants are unknown and the finding of their difference is the object of this problem.

4. The potential difference between the electrodes is measured by a balancing method or by means of a device with an infinitely large input impedance and with a total current in the boundary between the electrodes and the liquid equal to zero:

$$\int_{-\frac{b}{2}}^{+\frac{b}{2}} \frac{\partial \varphi}{\partial x}\Big|_{x=0} dy = \int_{-\frac{b}{2}}^{+\frac{b}{2}} \frac{\partial \varphi}{\partial x}\Big|_{x=a} dy = 0.$$

From Green's formula for any two functions F_1 and F_2 we have

$$\iint_S (F_1 \Delta F_2 - F_2 \Delta F_1)\, ds = \oint \left(F_1 \frac{\partial F_2}{\partial n} - F_2 \frac{\partial F_1}{\partial n} \right) dl. \tag{2}$$

Let $\Psi(x, y)$ be the solution of (1) which satisfies the boundary conditions of the problem. Assuming that $F_1 = \Psi(x, y)$ and $F_2 = 1$, we have

$$\oint \frac{\partial \Psi(x, y)}{\partial n}\, dl = \iint_S \Delta \Psi(x, y)\, dx dy = -\frac{\mu H}{c} \iint_S \frac{dv_z(x, y)}{dx}\, dx dy. \tag{3}$$

Let us examine the range $0 \leq \xi \leq x$; $0 \leq y \leq b$.

According to the second and fourth boundary conditions the integral round the contour is equal to

$$\oint \frac{\partial \Psi(\xi, y)}{\partial n}\, dl = \int_0^b \frac{\partial \Psi(\xi, y)}{\partial \xi}\Big|_{\xi=x} dy = \int_0^b \frac{\partial \Psi(x, y)}{\partial x}\, dy. \tag{4}$$

The area integral is equal to

$$\iint_S \frac{dv_z(\xi, y)}{\partial \xi}\, ds = \int_0^x \int_0^b \frac{\partial v(\xi, y)}{\partial \xi}\, d\xi dy = \int_0^b [v_z(\xi, y)|_{\xi=y} - v_z(\xi, y)|_{\xi=0} + \psi_1(y)]\, dy, \tag{5}$$

where $\psi_1(y)$ is a certain arbitrary function.

According to the first boundary condition

$$v_x(\xi, y)\big|_{\xi=0}=0$$

and from the expressions (3-5) it follows that

$$\int_0^b \frac{\partial \Psi(x, y)}{dx}\, dy=-\frac{\mu H}{c}\int_0^b [v_x(x, y)+\psi_1(y)]\, dy. \tag{6}$$

Multiplying (6) by dx and integrating between 0 and a we obtain

$$\int_0^b [\Psi(x, y)\big|_{x=a}-\Psi(x,y)\big|_{x=0}+\psi_2(y)]\, dy=\frac{\mu H}{c}\int_0^a\int_0^b [v_x(x, y)+\psi_1(y)]\, dx dy, \tag{7}$$

where $\psi_2(y)$ is a certain arbitrary function.

According to the third boundary condition

$$(\Phi_2-\Phi_1)b=\frac{\mu H}{c}\int_0^a\int_0^b v_x(x, y)\, dx dy+\text{const}. \tag{8}$$

For a velocity of the liquid equal to zero the difference of potentials between the electrodes is also equal to zero, and the arbitrary constant in the right-hand side of the equation disappears.

It follows from (8) that

$$u=\Phi_2-\Phi_1=\frac{\mu H}{c}\bar{v}a=\frac{\mu H}{c}\cdot\frac{Q}{b}, \tag{9}$$

where $\bar{v}=\frac{1}{ab}\int_0^a\int_0^b v_x(x, y)\, dx dy$ is the mean velocity.

Thus the readings of electromagnetic flowmeters with rectangular channels and electrodes which equalize their electric field will not depend on the distribution curve of velocities providing the channel is long enough and the magnetic field sufficiently uniform.

In an actual instrument the magnetic field is not uniform. An analysis of these conditions showed that for a nonuniformity of the field not exceeding 1% the effect of the velocity-distribution curve on the instrument readings is negligible.

Since the dimensions of the instrument along the flow are limited there will be a three-dimensional distribution of potential. Three-dimensional effects may arise due to the velocity, magnetic field or the electrical field gradient in the direction of the flow. The velocity gradient due to the transition from a pipe with a round cross section to the rectangular instrument channel occurs outside the magnet and should not influence the distribution of potentials. The magnetic field gradient (at the ends of the magnet) produces circulating currents in a plane perpendicular to the magnetic field.

Analysis has shown that the drop of potential between the electrodes due to circulation currents is determined both by the longitudinal gradient of the magnetic field and by the transverse velocity gradient, and depends on the distribution of velocities. The potential difference in this case is equal to

$$u = \frac{\mu H}{c} \bar{v} a \left(1 - \frac{\sum_{n=1}^{\infty} \frac{v_n}{n\pi} e^{-\frac{n\pi l}{2a}}}{\sum_{n=1}^{\infty} \frac{v_n}{n\pi}} \right) = K_1 \frac{\mu H}{c} \bar{v} a, \tag{10}$$

where v_n is the coefficient of the velocity distribution curve expansion in a Fourier series ($n = 1, 3, 5 \ldots$); l is the length of the magnetic field along the flow.

The values of coefficient K_1 are shown in Fig. 3 for the two extreme cases, a laminar flow and a turbulent flow for $Re \to \infty$, when the variation of velocity occurs in a very thin boundary layer and the velocity distribution curve with a certain approximation can be considered to be flat.

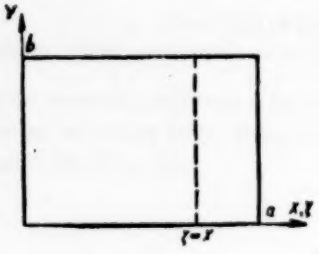

Fig. 2.

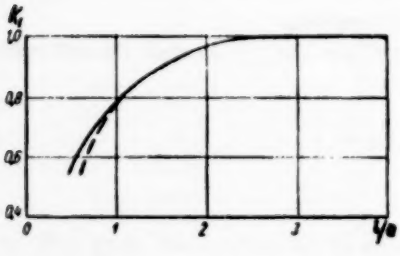

Fig. 3. —— laminar flow; – – – turbulent flow for $Re \to \infty$.

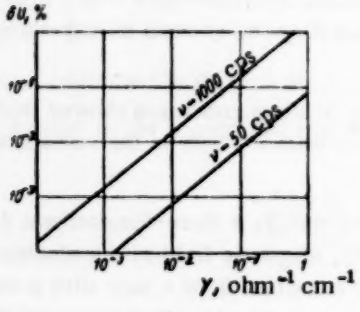

Fig. 4.

In order to prevent external interference the screens are usually grounded to the pipe. The potentials near the end of the channel, which is made of an insulating material, are equal to zero and produce a longitudinal gradient in the electric field. It has been established by the author of this article that the difference of potentials between the electrodes in this case also depends on the velocity distribution curve and is equal to

$$u = \frac{\mu H}{c} \bar{v} a \left(1 - \frac{\sum_{n=1}^{\infty} \frac{v_n}{n\pi \operatorname{ch} \frac{n\pi l}{2a}}}{\sum_{n=1}^{\infty} \frac{v_n}{n\pi}} \right) = K_2 \frac{\mu H}{c} \bar{v} a, \qquad n = 1, 3, 5 \ldots \tag{11}$$

where l is the length of the instrument channel insulated section.

The end effects connected with the three-dimensional distribution of potentials are reduced with an increased length of the instrument. If the length of the magnetic field along the flow and the length of the insulated section of the channel are 3-3.5 times greater than the width (diameter) of the channel, the end effects become independent and are no longer coupled with the velocity distribution curve effect.

The currents flowing in the instrument channel produce an indirect relationship between the readings and the physical properties of the measured liquid. For a sufficiently high conductivity of the liquid, the force of electromagnetic interaction between the currents and the magnetic field becomes comparable to the friction force and affects the movement of the liquid, thus disturbing the symmetry of the flow. Since the readings of an electromagnetic flowmeter with a rectangular channel do not depend on the distribution of velocities, the magnetohydrodynamic disturbances should not affect the readings.

The degree to which the magnetohydrodynamic disturbances affect the readings of the flowmeter with a round channel is determined by the square of Hartmann's number [6] and depends on the magnetic field strength, channel dimensions and the relation between the conductivity and viscosity of the liquid. The errors become noticeable at conductivities exceeding 0.1 $ohm^{-1}\ cm^{-1}$.

Flowmeters designed for liquids with an ionic conductivity usually operate with an alternating magnetic field $H = H_0 \sin\omega t$. The electric field in the channel of such instruments usually consists of a superposed potential and solenoidal fields. The potential field is determined by the movement of the liquid, but the solenoidal does not depend on velocity. The effect of the solenoidal field on the meter readings is two-fold. It introduces noise voltages into the measuring circuits. These disturbances do not depend on the physical properties of the liquid and can be compensated. The secondary magnetic fields of eddy currents in the measured liquid distort the instrument's magnetic field, disturbing its uniformity. Inphase errors connected with this effect, and found by the author of this article, are shown in Fig. 4.

When flows of liquids with a low conductivity are measured the effect of displacement currents becomes noticeable. The displacement currents complete the circuit through the surrounding space, thus producing in the area between the electrodes a drop in potential, which depends on the relation between the dielectric constants of the liquid and the surrounding media. The effect of the displacement currents becomes noticeable for conductivities of the order of 10^{-8} to 10^{-9} ohm^{-1} cm^{-1}.

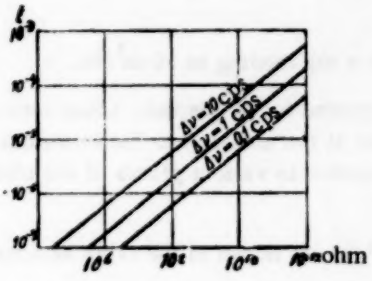

Fig. 5.

The resistance of the liquid between the electrodes is a source of thermal noise, which limits the sensitivity of the flowmeter. The noise level is determined to a great extent by the frequency characteristic of the measuring device, and can be reduced by using narrow-band devices with a large time constant. Figure 5 shows the relation of the noise level to the internal resistance of the flowmeter for different frequency bandwidths. For a time constant of several seconds and a signal level of 10^{-3} to 10^{-2} the effect of thermal noise becomes noticeable if the resistance of the liquid between the electrodes is of the order of 10^8 ohm (conductivity to 10^{-7} to 10^{-8} ohm^{-1} cm^{-1}).

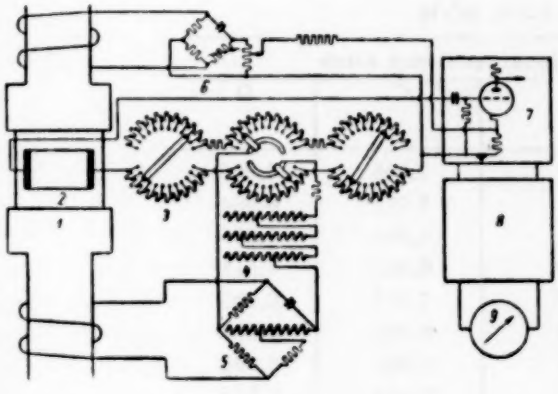

Fig. 6.

Thus the influence of the physical properties of liquids appears in the form of secondary and stray effects when the conductivity of the liquid is greater than 10^{-1} ohm^{-1} cm^{-1} or less than 10^{-7} to 10^{-8} ohm^{-1}. cm^{-1}, i.e., electrical flowmeters can be used with the majority of commercial liquids. In the experiments conducted by the author of this article the effect of Reynolds' number and local resistances on the instrument reading was considered separately.

Investigation of the effect of the velocity distribution curve was carried out over a wide range of Reynolds numbers, from 900 to 500,000. * For Reynolds numbers between 900 and 50,000 a flowmeter with a top measuring limit of 10 m^3/hr (Fig. 6) was used, and for numbers 50,000 to 500,000 a flowmeter with a top reading of 500 m^3/hr.

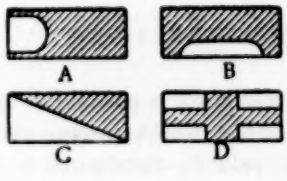

Fig. 7.

For measuring the signal which arises between the electrodes of the flowmeter transducer chamber 2 a Disselhorst three-decade low-resistance potentiometer 3 was used. The operating circuit of the potentiometer was fed from a coil fixed to the core of electromagnet 1. The operating current was set during the graduation of the instrument by a known flow by means of the four-decade resistance box 4. The phase of this current was

*Reynolds numbers are referred to the hydraulic radius of the instrument channels.

adjusted with phase-shifter 5. In order to protect it from external interference the potentiometer was placed in a double iron screen. The imbalance signal was amplified in a narrow-band amplifier 7 and detected by a synchronous detector 8. A moving coil microammeter 9 was used as a zero indicator. The zero of the instrument was set by means of zero-adjuster 6. The flowmeter worked at a frequency of 23 cps (a flow of 10 m^3/hr corresponded to a signal of the order of 2 mv).

The cross section of the instrument channel was 12 × 30 mm. The length of the rectangular part of the channel was 260 mm. The dimensions of the pole pieces were 200 × 200 mm, and the gap between them 30 mm. The nonuniformity of the magnetic field in the working portion was less than 0.1%. The ratio of the length of the uniform part of the magnetic field to the width of the channel exceeded 3.5.

Reynolds' critical number was determined experimentally by introducing into the flow some tinted liquid.

A flowmeter with a top reading of 500 m^3/hr worked at the commercial frequency. The signal was compensated by means of a slidewire. The compensator-operating circuit was fed from a special coil fixed to the core of the magnet. The instrument was automatically balanced by a reversible capacitor-starting motor.

Test results have shown that the flow coefficient does not depend on Reynolds' number and hence not on the velocity distribution curve.

The effect of local resistance was investigated on a flowmeter with a top reading of 10 m^3/hr.

The transition from a circular pipe to a rectangular channel was provided by an adaptor. Local resistance was provided by vanes, which were inserted on the near side of the adaptor at the entrance to the instrument channel. The shape of the vanes was made to introduce considerable distortion in various planes of the channel (Fig. 7).

The series of measurements made at the instrument reading of 7 m^3/hr are shown in the table attached.

A statistical analysis of the measurement results showed that their dispersion is random and does not depend on the distortions of the flow.

Flow, m^3/hr				
Without a vane; channel completely open	Shape of local resistance vanes			
	A	B	C	D
6.991	7.011	6.985	7.005	6.990
6.999	6.983	7.000	7.011	6.984
7.007	6.975	6.974	6.994	7.009
6.997	6.981	7.022	6.988	7.011
7.011	7.000	6.982	7.018	6.989
6.984	7.001	6.979	6.993	6.982
6.992	6.988	7.008	6.986	6.998
7.006	7.007	7.005	7.018	6.995
7.004	6.999	6.988	6.985	6.983
6.991	6.988	7.000	6.997	6.990
Average				
6.988 ± 0.007	6.993 ± 0.009	6.994 ± 0.011	6.999 ± 0.009	6.993 ± 0.006

The experimental results have confirmed the theoretical considerations and make it possible to assert that the readings of electromagnetic flowmeters with rectangular channels and electrodes which average out the electric field do not depend on the velocity distribution curve for any arbitrary velocity distribution in the instrument channel.

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INFORMATION

FIRST INTERNATIONAL CONGRESS OF THE IFAC ON AUTOMATIC CONTROL

I. E. Dekabrun

Translated from Izmeritel'naya Tekhnika, No. 10, pp. 60-63, October, 1960

The first international congress of the IFAC on automatic control was held in Moscow from June 27 to July 2, 1960. The congress, which was organized by the national committee on automatic control of the Soviet Union, was attended by 1,111 delegates from 29 countries and over 1,000 guests. Two plenary and several sectional sessions were held, at which 274 papers were presented and over 580 contributions to the discussion were made.

The work of the congress, which proceeded under the slogan "Widely applied theory, most reliable equipment, highly efficient automatic systems," was divided into three sections: automatic control theory, equipment for automation and remote control, and their application.

Both the papers themselves and the discussion on them served as a vivid demonstration of the high level of technical development in this sphere, based on the wide application of mathematical methods and devices.

In the sphere of automatic equipment the congress dealt with electrical and magnetic control system elements; electrical calculating and simulating devices, programming elements and machines: transducers, components and systems of remote control, pneumatic equipment used in automation and calculating machines, automatic control devices and systems. Papers dealt with methods of designing and applying various elements and devices, as well as with methods of electrical measurements.

The papers read at the congress and the discussion on them showed that the basic trends of development in automatic equipment consist of the following.

The finding and development of new principles in converting parameters of automatic control and regulation systems, including the development of new methods of automatic measurements of parameters and of the automation of existing measuring methods.

The finding and development of new principles for obtaining, processing, transmitting and using information in automatic control and regulation systems. Transition from an individual control of parameters to centralized systems of collecting and processing information. A wide use of digital techniques for presenting and processing monitoring and control information.

Finding and developing most effective principles for transmitting information in automatic and remote control systems.

Finding and developing new methods and techniques in constructing equipment and raising its reliability, accuracy and life. Transition from electromechanical to contactless elements based on the use of magnetic and semiconductor components. Development of new principles in designing pneumatic and hydraulic equipment based on the utilization of jet interaction.

Development of new technology (printed circuits, miniature components, etc.) Technological problems in the choice and production of equipment.

Below we give a brief resumé of some of the papers read on measurement techniques.

V. S. Malov, A. M. Pshenichnikov and Ya. A. Kupershmidt (USSR) "Commercial Remote Measuring Systems and Digital Techniques." A classification of remote measuring systems is given on the basis of the following criteria: 1) the distance of transmission and hence the type of communication channels; 2) the construction of the remote measuring system; 3) the type of remote measuring parameters (electrical and nonelectrical); 4) operating time. The possibility and desirability of unifying remote measuring systems and using unit construction are indicated. It is shown that the application of digital techniques in remote measuring systems has great possibilities, since it will lead to the extension of the scope of remote control systems. The paper cites examples of digital reproduction systems: with the transmission of coded or analogue signals.

V. L. Inosov and V. A. Lutskii (USSR), "Coded-Pulse Remote Measuring System for Controlling Main Gas Pipe-Lines." This system and its main units are described in detail. The system is made up of contactless elements: semiconductor and ferrite components with a minimum utilization of nonstandard units or details. A coded-pulse method of discrete remote measurements is used, thus providing a high degree of resistance to noise. Recording is made by digital registration on a printing machine. The combination of this registration with a coded-pulse transmission method provides high accuracy, which is only limited by the errors in the transmitter. The conversion of manometer and flowmeter readings into a discrete form is made by means of a space encoder with a mask and photo-cells but without any intermediate devices. In order to avoid read-out errors Gray's mask is used with a binary code. For manometers it is linear, for flowmeters it is proportional to the square root of the rotation angle of the mask.

The system requires three transmitting frequencies: two for transmitting the quality 0 and 1 and one (the third) for alarm and auxiliary signalling. Monitoring and transmission of measurements from control points (a maximum of 30) to the central control station are provided. The largest amount of data transmitted from one point amounts to 8 remote measurements and one alarm signal. The system can operate over telephone carrier channels or two-wire lines.

V. N. Mikhailovskii (USSR), "Evaluation Criteria and Possibilities for Raising the Quality of Remote Measuring Systems." The application of the basic propositions of the information theory to solving various remote measurement problems is examined. Information theory criteria for specific remote-measurement systems are cited. It is shown that the quality of a remote-measurement system is most completely and objectively evaluated by information criteria, since they provide comparative evaluation of remote-measuring systems both on the basis of their potentialities apart from the concrete conditions of their use, and on the basis of the efficiency of their employment under given conditions. By analyzing expressions containing information criteria it is possible to find ways of raising the quality of remote measurements and remote-measuring systems.

S. S. Dreiper, M. Finstone, V. G. Denhard, and D. Goldenberg (USA), "Dynamic Suspension of Instrument Components by Means of Viscous Liquids." Hydrodynamic suspension on a float of basic moving elements of instruments provides high accuracy of operation under conditions of shocks, vibrations and acceleration. Correctly selected gaps and viscosity of liquids may almost completely eliminate the effect of friction in the supports. Under dynamic conditions the suspended element is held in a given position by the redistribution of pressure due to the flow of the liquid in the gap. A general theory of hydrodynamic suspension was worked out; the hydrodynamic suspension of a cylindrical moving element subjected to various acceleration and vibration effects was examined; a comparison with an ordinary suspension was made; certain experimental results were reported.

S. I. Li (USA), "Servovalve with a Discharge Feedback and Dynamic Characteristics of a System Consisting of a Pilot Valve and an Inert Load." The paper deals with quick-acting flowmeters of the throttle-plunger type and special mechanical devices for adding up signals, thus making it possible to use two such flowmeters for measuring flows which are rapidly changing both in magnitude and direction. These flowmeters provide an accurate control of a servovalve output by measuring this output and using the measured value as a feedback signal. This reduces considerably the accuracy required in manufacturing pilot valves which control the flow. An account is given of the tests made on a system consisting of a pilot valve with a discharge feedback and a comparison made with a system containing an ordinary servovalve.

A. Tastin (Britain), "Thermal Operating Conditions of Automatic Control System Components." The paper deals with problems of calculating or selecting automatic control system components for the purpose of eliminating excessive temperature when working under given conditions. It is shown that if with a given design excessive temperatures are not attained, it is possible to improve the control characteristics of the system by raising

its loading. Certain methods of determining the heating of components in any part of the system subjected to a definite varying load, which can only be determined statistically, are described. The principal method is based on "potential temperatures," which correspond to various operational conditions. It is also shown that in certain cases simpler methods can be used, which consist of expanding variables in Fourier series and making use of quadratic mean evaluations.

E. K. Lloyd and D. P. Johnson (USA), "Static and Dynamic Calibration of Instruments for Measuring Pressure by the National Bureau of Standards." The paper describes an installation for calibrating pressure-measuring instruments developed and used by the National Bureau of Standards (USA). The equipment for static calibration of instruments includes an oil-piston manometer for high pressures with an adjustable gap, an air-piston manometer (normal and sloped), and precision mercury manometers. For dynamic calibration at high pressure a pressure-pulse generator is used which provides precision amplitude values, and for lower pressures a shock-tube. The sphere of application of these instruments, their relative accuracy and the source of errors are given. The contemporary state of this sphere of metrology is examined from the point of view of the accuracy of pressure standards and certain projects for improving measuring methods at high pressures are given.

U. H. Leslie (Britain), "Precision System for Measuring Torque by Means of a Balanced Machine." A digital system for controlling the speed of rotation applicable to any type of motor is described. The speed of the shaft rotation is converted into a train of pulses whose frequency corresponds to the speed of rotation of the shaft. This frequency is compared in a digital differential device with that of a crystal quartz oscillator. The output signal of the differential device is used for controlling the motor supply circuit and adjusting the speed of its rotation. A system for measuring the torque of hydraulic turbines and pumps is described. The body of the balance machine is supported on frictionless oil hydrostatic bearings. The torque is measured by means of a hydraulic servomotor which converts the moment of resistance to rotation, which is applied to a lever with a known arm, into oil pressure. This pressure is measured by a manometer and can be recorded. The error in measuring the torque amounts to 0.01%.

A. Nomoto, T. Koyke, T. Yamaguchi, and H. Kawamoto (Japan), "Automation of the Measurements of Dynamic Characteristics." Two different types of instruments for measuring the frequency characteristics of control systems are described. The operation of the instruments is based on the same principle, which consists in determining the correction required to make the measured output signal correspond with a reference sinusoidal signal. The basic component of the measured signal is separated at the output in order to determine the transfer function of both the linear and nonlinear systems. Both instruments consist of a variable frequency signal generator, a receiving device and a device for processing data. Calculations are made in an analogue form.

In the instrument of the first type correlation is determined by means of an active integrating circuit and the result is recorded and visually displayed in the form of an amplitude-phase characteristic. It covers a range of 0.02-100 rad/sec and serves to measure production-control systems.

The instrument of the second type uses a compensation method. The phase of a reference sinusoidal signal is automatically compared with that of the measured signal. The gain is measured by an automatic potentiometer. By means of this instrument amplitude-phase or amplitude-frequency and phase-frequency characteristics can be obtained directly. Instead of integrators passive filters are used for determining correlation. Its range is 0.6-300 rad/sec. The instrument is used for investigating monitoring systems.

J. Vidal (Belgium), "Automatic Data-Processing System for Thermal Investigation Chambers." The paper describes a system for measuring and processing data of a test chamber with calorimetric walls, designed for measuring heat transmission into the ambient space, the effect of heating obtained from external sources, and heat losses in constructional panels and materials. The automatic system provides for recording a large number of variables according to a definite program. The monitored parameters are registered by converting analogue signals into a digital form. Automatic processing of data is made by means of computer IBM-650 and auxiliary equipment in the calculating center.

D. A. Patient (Britain), "Automation of Selection and Chemical Analysis of Samples." The paper describes methods and equipment for automatic checking of the composition of substances in the basic method of selecting samples in the chemical analysis of liquids. A detailed description is given of the design of devices used for automatic control of the flow of liquids (electromagnetic and pneumatic valves), for the selection of samples and

the measurement of the volume of liquids. Attention is paid to the problem of reliability and simplicity in use. Means for recording the measurement results are examined, as well as methods for transmitting them to a control system. Examples of the application of this instrument are given. Great attention is paid to the automatic apparatus for measuring and checking the properties of liquids. Two methods of processing the samples obtained from the production installations are described, those of direct reaction and extraction. Special attention is paid to the description of the method for complete automation of titration, the most important method of analysis.

D. I. Ageikin (USSR), "Compensated Thermomagnetic Gas Analyzers." In the existing thermomagnetic gas analyzers for analyzing oxygen, the intensity of the thermomagnetic convection is measured by thermoanemometric methods, by determining the cooling it produces in the operating heaters. The possibility of measuring the drop in pressure, caused by thermomagnetic convection, by means of compensating it with an artificially produced pressure of the opposite sign is discussed. Schematics of such gas analyzers were given and their basic characteristics discussed.

S. D. Schechter and F. N. Muntianu (Rumania), "Comparative Evaluation of Circuits for Automatic Zero Stabilization in DC Operational Amplifiers." The paper examines the basic errors of dc operational amplifiers; it shows that the most important one of them is that due to zero drift. A comparative evaluation of various circuits decreasing the zero drift is given. The circuits are compared both from the technical and economic point of view.

K. A. Netrebenko (USSR), "Design Principles of Digital Binary-Coding Automatic Compensators." Certain new principles in the design of the units of digital autocompensators are proposed. The paper examines certain new circuits of parallel potential dividers for binary-coding automatic potentiometers, and shows their advantages as compared with ordinary circuits, consisting in a decreased number of resistances, simplicity of switching, less rigorous requirements for the quality of switches, possibility of using contactless switching, etc. In the case of digital autocompensators which work by the method of order by order balancing, means are indicated for speeding up their operation and simplifying the design of the pulse-distributor. The use of intermittent-acting amplifiers with a periodic zero-adjustment is recommended in zero elements of digital autocompensators.

K. B. Karandeev and G. A. Shtamberger (USSR), "Quasi-Balanced Bridge as an Element of an Automatic Control System." A very convenient element for the solution of many conversion problems, obtaining information, etc., consists of an impedance, its components and derived parameters. The paper describes methods for continuous measurement and registration of variations in information parameters, which occur fairly rapidly. The proposed method of using the ac bridge for scaler measurements simplifies the measuring and control process. In this connection two versions are possible, adjusting the circuit to a modulus or phase-measuring condition. Both these conditions provide single-valued results. Analytical conditions characterizing the separate measurements of components for either condition are given. Specific versions of the quasi-balanced bridges for separate measurements of the real and imaginary components of impedances are described.

PAPERS READ ON THE TECHNIQUE OF RADIO MEASUREMENTS AT THE ALL-UNION SCIENTIFIC SESSION DEDICATED TO THE RADIO DAY IN 1960

Translated from Izmeritel'naya Tekhnika, No. 10, pp. 63-64, October, 1960

The All-Union Scientific Session dedicated to Radio Day was held in Moscow from May 16 to 20, 1960, and was divided into 16 sections including a section for radio measurements.

At the four sessions of the Radio Measurement Section fourteen papers were read and discussed.

The first session dealt with the problems of frequency stabilization.

The paper of A. Ya. Leikin and E. Z. Orlov on "Coincidence in the Frequencies of Molecular Generators of Similar Construction" dealt with the investigation of the coincidence in frequencies of separate molecular generators of similar construction, tuned by the same method under definite conditions. The possibility of transferring the value of a frequency produced by a molecular generator under prolonged investigation to newly-made generators was demonstrated.

L. D. Bryzzhev's paper entitled "Quartz-Crystal Generator with Automatic Phase-Trimming"* dealt with methods of improving the stability of quartz-crystal generators by means of phase-trimming. In the proposed circuit frequency deviations of the crystal generator due to any factors produce an error signal in the form of a voltage of the main frequency displaced in phase by 90° with respect to the input voltage. This signal is then used for reducing the frequency deviation. The method was checked experimentally on a semiconductor quartz-crystal generator with satisfactory results.

In his paper on "100 kc Quartz Resonators with a Q Factor of $(10\text{-}15)\cdot 10^6$," I. V. Grinenko dealt with the work carried out for the improvement in the design and the parameters of a quartz-crystal resonator. As the result of improving the quality of the surface of the vibrating crystal, a better design of the crystal holder and improved technology in making the quartz elements, 100 kc quartz-resonators were made with a Q factor of 10-15 millions.

E. D. Novgorodov shows in his paper entitled "Relation of the Short-Term Instability of a Quartz Oscillator to the Q Factor of the Quartz Resonator" that the short-term instability of a quartz-oscillator frequency is mainly due to the short effect and is inversely proportional to the square root of the resonator Q factor. It was assumed that the theoretical conclusions on the relations between the shot effect and phase distortion obtained for a tube-oscillator with an ordinary tuned circuit could be extended to a quartz crystal oscillator. Experimental determination of second-long instability in a quartz oscillator for Q factors of the resonator in the range of 3.6 thousands of 4.3 millions was in good agreement with theoretical conclusions.

In his paper P. A. Shpan'on described a method of measuring frequency deviation in frequency modulation by means of frequency counters.

As a result of the discussions of the papers dealing with the problem of frequency stabilization the section noted the outstanding interest and practical value of the results obtained.

At the second session of the section papers were read by M. V. Fomin on "Panoramic Waveguide Reflectometer for Radio Relay Lines," and by the Hungarian expert A. T. Sarkany on "Highly-Sensitive Wide-Band Tube Voltmeter for Measuring Effective Voltages."

The section noted the importance of the instrument described in M. V. Fomin's paper.

The highly-sensitive wide-band tube voltmeter described by A. T. Sarkany can be widely used for improving the technique of measuring the parameters of radio communication lines and in laboratory practice by using it as a highly-sensitive voltmeter.

At the third and fourth sessions of the section seven papers were read dealing with various methods of radio measurements and radio-measuring instruments.

After listening to V. P. Kovalev's paper entitled "Experimental Investigations of the Electromagnetic Field at UH Frequencies and Peculiarities of the Measuring Apparatus," the session noted the scientific and technical value of this method of investigating the amplitude and phase distributions of the UHF electric field by means of a passive-modulated probe with the possibility of a simultaneous and automatic measurement of the field components. The section recommended this paper to be published.

With respect to A. I. El'kind's paper on "Coaxial Slotless Measuring Lines,"** the section noted the scientific and practical value of the expounded theory and the design of slotless measuring lines, which have a greater accuracy than the commonly used slotted lines. The section recommended to adopt the slotless lines for measuring purposes.

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*Izmeritel'naya Tekh. 8, 30 (1960). [See English translation.]

**Izmeritel'naya Tekh. 9, 44 (1960). [See English translation.]

Having heard K. N. Shchelkunov's paper on the development of an instrument for measuring secondary radiation by means of a reflecting klystron-autodyne, the section noted the usefulness of this work for investigating the reflecting properties of various materials. The paper was recommended for publication.

With respect to L. P. Kolesnikov's paper on "Automatic Standing-Wave Ratiometers" the section noted their practical value and recommended their use in measurements.

In A. L. Grokhol'skii's paper on "Standards of Conductance, Susceptance and Small Capacitance," he dealt with reference measures and apparatus used for checking instruments which measure parameters of circuits and materials. The section considered the work of L. P. Kolesnikov and A. L. Grokhol'skii valuable and deserving wide publicity.

With respect to the papers of S. I. Borovitskii on "Set of Instruments for Measuring the Noise Factor" and G. P. Vikhrov on "Reference Digital Measuring Apparatus for Checking Pulse-Generators," the section noted their practical value.

METROLOGICAL SECTION BUREAU OF THE CENTRAL ADMINISTRATION OF THE INSTRUMENT-MAKING INDUSTRY'S SCIENTIFIC AND TECHNICAL SOCIETY

Translated from Izmeritel'naya Tekhnika, No. 10, p. 64, October, 1960

A session of the Metrological Section Bureau of the Central Administration of the Instrument-Making Scientific and Technical Society and the All-Union Scientific Research Institute of the Committee of Standards, Measures and Measuring Instruments, together with representatives from plants and primary organizations of the instrument-making industry's scientific and technical society was held on July 12, 1960.

A. P. Bukarev noted that measurements play an increasing part in instrument-making and that development of the theoretical foundations of production control is of primary importance together with the solution of the fundamental theoretical problems connected with the assimilation of modern types, methods and means of measurement.

It is necessary to develop the foundations of production control in automatic manufacturing processes. Many industrial laboratories limit their activities to checking equipment and do not pay sufficient attention to the assimilation of new measurement techniques, equipment and methods in connection with total mechanization and automation, the introduction of conveyor-belt and automatic lines and the replacement of old equipment.

S. S. Podlazov in his paper on the basic trends in the development of automatic control dealt in detail with the principal tasks in the sphere of the application of the control and the strict requirements imposed by science and technology on the production of the control equipment. In many instances the error of measurement must not exceed 0.1-1μ with a speed of operation of 1-1.5 sec. The checking equipment must be stable for tens of hours and its life must equal 30,000-90,000 hours.

The paper paid special attention to automatic control equipment in automatic lines, in which various transducers are used as controlling devices.

It is necessary to organize the centralized manufacture of transducers, reduce their size, and raise their accuracy and reliability.

The speaker noted the possibility of using memory devices in the control equipment and the necessity of measuring many parameters.

Ya. M. Git's paper dealt with electrical contact transducers for controlling linear quantities and displacements, and their use for automatic and mechanized control purposes. He enumerated the basic requirements of the transducers.

The speaker noted the advantages and increased reliability of transducers made with gold-nickel or platinum-iridium contacts.

The Metrological Section Bureau decided to hold in October-November, 1960, a scientific and technical conference dealing with modern achievements in the sphere of automation and mechanization of control operations in linear and angle measurements, and with the assimilation of advanced measuring equipment and methods in instrument making.

A. B.

COMMITTEE OF STANDARDS, MEASURES AND MEASURING INSTRUMENTS

NEW SPECIFICATIONS FOR MEASURES AND MEASURING INSTRUMENTS APPROVED BY THE COMMITTEE

NEW STANDARDS

(Registered in July-August, 1960)
Translated from Izmeritel'naya Tekhnika, No. 10, p. 64, October, 1960

GOST (State Standard) 3059-60. Reference levels. Replacing GOST 3059-53.

GOST 5944-60. Rack-Mounted Measuring Instruments. Bodies. Over-all dimensions. Replacing GOST 5944-57.

GOST 9483-60. Weighing Instruments. Clock-type, square and round dials. First issue.

GOST 9486-60. AC Measuring Bridges. First issue.

GOST 9500-60. Portable reference dynamometers. First issue.

GOST 9504-60. Feeler instruments for determining surface roughness. Types. Basic parameters and accuracy standards. First issue.

GOST 9509-60. Weighing instruments. Knife-edge bearings and pads. First issue.